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RECENT DEVELOPMENTS IN ELECTRONIC-FLASH LAMPS

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Photography with electronic-flash lamps has become such a commonplace in recent years, amongst amateurs as well as professionals, that these lamps now rival in importance the older combustion-type flash bulbs. This is evident from the fact that the annual world production of electronic-flash units has been roughly half a million over the past five years. Assuming that each outfit is used on the average, say, 40 times a year, we see that a total of 100 000 000 photographs will be made this year with electronic-flash lamps. A development on this scale has been made possible by the progress achieved in the quality, economy, and reliability of the flash lamps themselves and the apparatus in which they are used. The principles underlying the design and manufacture of modern electronic-flash lamps are dealt with in the following article.

Introduction

Since the era of the explosive, smoky flash powders, the technique of flash photography has developed in two directions. In the first place there are the combustion-type flash bulbs, that can only be used once [1]); the flash unit for these bulbs contains a minimum of circuitry, namely a battery, a resistor and a capacitor. Secondly, there is the electronic-flash lamp, which was a later arrival on the scene. Originally, this type of lamp was only intended for special purposes, but subsequent development has now given it a firm footing in many fields of photography, including amateur photography. The circuit for the operation of the electronic-flash lamp is rather more elaborate than that needed for the combustion flash bulb, but it has the advantage of enabling large numbers of shots to be taken in fairly rapid succession. An article on the development of electronic-flash lamps was published in this journal seven years ago [2]).

Progress since that time has been marked by the advent of lamps for operation at lower voltages, making it possible to replace the paper capacitors originally used in the flash units by electrolytic capacitors, which are much lighter and smaller, but are not suitable for voltages higher than about 600 V.

The requirements of the designer of flash units have to be taken into account in the design of electronic-flash lamps, not only in the question of operating voltage but quite generally. This has led to the wide variety in the shape and dimensions of the electronic-flash lamps nowadays being made. As an illustration of this *fig. 1* shows a selection from the range of types manufactured by Philips.

According to their application the numerous types of electronic-flash lamps can be divided into four categories: *a*) flash lamps for voltages between 400 and 500 V, used in outfits with electrolytic capacitors; *b*) flash lamps for voltages between 2000 and 3000 V, used in outfits with paper capacitors; *c*) flash lamps for special scientific and industrial applications; *d*) stroboscopic lamps. In this article we shall be concerned only with flash lamps of categories *a*) and *b*), intended for general use.

Some remarks may be made here about the other two categories. The electronic-flash lamps under *c*), for special scientific and industrial purposes, closely resemble in many respects the types for general use. Apart from the shape of

*) Lighting Division, Eindhoven.

1) J. A. M. van Liempt and J. A. de Vriend, The "Photoflux", a light-source for flashlight photography, Philips tech. Rev. **1**, 289-294, 1936.
L. H. Verbeek, The specific light output of "Photoflux" flash-bulbs, Philips tech. Rev. **15**, 317-321, 1953/54.
J. A. de Vriend, Ignition of "Photoflux" flash-bulbs with the aid of a capacitor, Philips tech. Rev. **16**, 333-336, 1954/55.

2) N. W. Robinson, Electronic flash-tubes, Philips tech. Rev. **16**, 13-23, 1954/55.
Instead of "electronic flash-tube" the term "electronic-flash lamp" as recommended by the C.I.E., will be used here; where no misunderstanding is possible, we shall often simply refer to flash lamp.

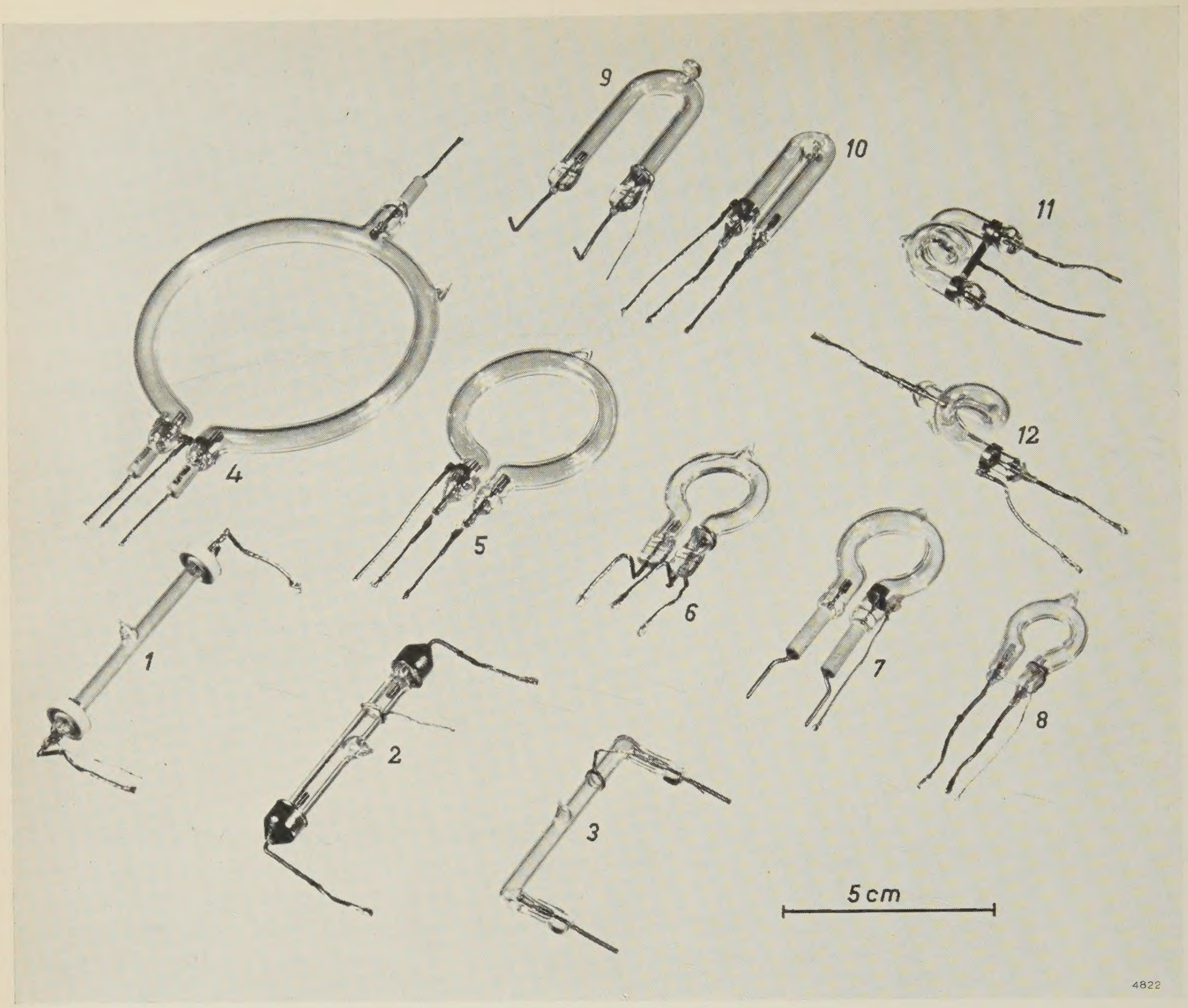

Fig. 1. Some examples of the wide variety of electronic-flash lamps made by Philips. Some types are U-shaped, some Ω-shaped, and others are in the form of spirals or straight tubes. All the types shown have flying leads. Their long life (more than 10 000 flashes) means in effect that they scarcely ever need to be replaced.

the lamps, the voltage, load and spectral distribution of the flash are adapted to the purposes for which the lamps are required. For example, there are special lamps for photographing tracks in Wilson cloud chambers [3]), for photographing the human eye [4]), for approach-warning on police cars, and for making photocopies.

The lamps in categories *a*), *b*) and *c*) are as a general rule suitable only for separate flashes or short successions of flashes, e.g. 10 to 20 flashes within a few minutes, after which they need some time to cool down. The stroboscopic lamps under *d*), however, are required to operate for up to several hours at a time, with a continuously variable flash frequency of anything from 15 to 300 flashes per second [5]). In the case of normal electronic-flash lamps, the permissible loading is given in watt.seconds per flash, while for stroboscopic lamps the average load in watts is usually specified, although of course these lamps, too, are pulse-loaded. It will be clear that the requirements imposed on stroboscopic lamps involve a whole range of fresh problems, with regard to both the design of the lamps and the circuitry.

In dealing with general-purpose electronic-flash lamps we shall first consider the electrical conditions under which the lamps have to operate. After then examining certain characteristics of the flash, we shall deal at somewhat greater length with design and manufacture. We shall then try to form a synthesis of all these questions in order to see how

[3]) N. Warmoltz and A. M. C. Helmer, A flash lamp for illuminating vapour tracks in the Wilson cloud chamber, Philips tech. Rev. **10,** 178-187, 1948/49.

[4]) J. E. Winkelman and N. Warmoltz, Photography of the eye with the aid of electronic flash-tubes, Philips tech. Rev. **15,** 342-346, 1953/54.
H. J. J. van Boort, N. Warmoltz and J. E. Winkelman, Colour photography of the retina and the anterior segment of the eye with the aid of a discharge flash lamp, Medicamundi **3,** 56-65, 1957.

[5]) S. L. de Bruin, An apparatus for stroboscopic observation, Philips tech. Rev. **8,** 25-32, 1946.

we may arrive at optimum designs of flash lamps. Finally, with reference to some typical representatives from the Philips range of flash lamps an idea will be given of the possibilities offered by this kind of lamp.

Electrical operating conditions of flash lamps

The energy to be dissipated in an electronic-flash lamp is derived from a capacitor, charged to a certain voltage. The energy stored in the charged capacitor is then converted into light in the flash lamp at the appropriate moment. This is done by discharging the capacitor through the lamp by means of a triggering circuit. There are thus two different electrical processes involved: the first process makes the required energy available, the second process, which is as a rule controlled from the camera, converts the energy into light at the right moment.

We shall now examine these two processes with reference to the circuit shown in *fig. 2*. The main capacitor, C_f, is charged to a suitable DC potential U; this potential also prevails between the anode A and the cathode K of the flash lamp B. In contrast to the arrangement described in reference [2]), the cathode is here earthed, that is to say connected to the frame of the apparatus; this has become increasingly the practice in recent years.

In parallel with the lamp is a voltage divider R_1-R_2, whose function is to reduce the voltage across the initiating contact S (in the camera) to 300 V, which is generally regarded as the maximum permissible [6]). The capacitor C_i and the primary of the trigger-pulse transformer T constitute the triggering circuit. When the contact S is closed, the discharge of the capacitor C_i induces a voltage surge

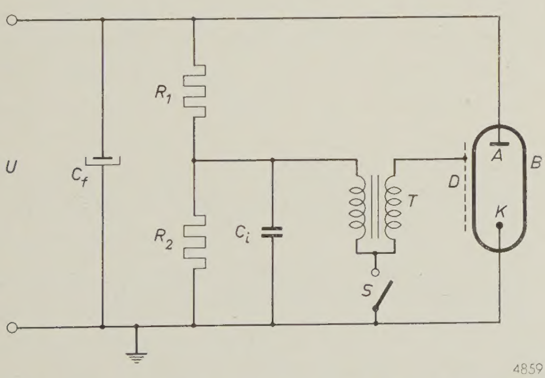

Fig. 2. Basic circuit for an electronic-flash lamp. The main capacitor C_f is charged to a DC potential U. B envelope, A anode, K cathode and D trigger electrode. R_1-R_2 voltage divider for the triggering voltage. C_i capacitor for trigger pulse. S initiating contact, normally built into the camera and synchronized with the shutter. T trigger-pulse transformer.

[6]) In Germany, where the most electronic-flash units for the European market are made, this value has been laid down in a standard recommendation (DIN 19 014).

in the secondary of the transformer T, and the trigger electrode D receives the high voltage pulse required to trigger the flash lamp.

The potential U can be obtained in various ways. Formerly, an accumulator or dry battery was generally used for this purpose in conjunction with an electromechanical vibrator, a transformer and a rectifier, but in recent years increasing use has been made of transistor circuits.

We shall now indicate the requirements to be met by the power supply, which are of importance in the considerations to follow. In order to produce the maximum number of effective flashes per dry battery or per charge of the accumulator, the load per flash must of course be as small as possible. Moreover the power supply must be so designed as to keep the final voltage which the main capacitor receives as constant as possible, since the energy accumulated in the capacitor varies according to the square of the operating voltage. A drop in the supply voltage U by $12\frac{1}{2}\%$ causes roughly a 25% drop in capacitor energy and hence in flash energy; the light output then drops by about the same percentage. This can have particularly unfortunate results in colour photography, owing to the relatively small latitude of colour-sensitive emulsions. From the photographic point of view, the above drop in light output means of course that the diaphragm ought to be opened a further half stop. Some improvement in this respect became possible with the advent of the control circuits used at the present time. These circuits meet both the above-mentioned requirements: they switch off the current source as soon as the capacitor is fully charged, and they automatically re-charge the capacitor as soon as the leakage current in the capacitor causes the voltage to drop below a certain preset value.

At a capacitance C and a voltage U the energy accumulated in the main capacitor is $\frac{1}{2}CU^2$. We shall disregard here the fact that the effective capacitance of electrolytic capacitors depends on the speed of the discharge. For an energy of, say, 125 Wsec, we then have the choice between a capacitance of, for example, 1000 μF at 500 V and a capacitance of 40 μF at 2500 V. In the latter case, in view of the high tension a paper capacitor must be used, which, on the given data, would weight about four pounds and have a volume of just over a litre; in the other case (lower voltage) we can use two electrolytic capacitors, having a total weight of about 2 pounds and a volume of about two thirds of a litre. A much smaller power pack is thus possible. Since it has become possible to make electrolytic capacitors capable of withstanding

repeated charge and discharge whilst retaining a constant capacitance, there has been a steady trend, in view of the advantages mentioned, towards flash lamps for lower voltages.

Further factors that militate in favour of lower voltages include safety measures (protection from high tensions becomes simpler) and the fact that for a given flash lamp and a given flash energy the duration of the flash can be longer. We shall return to this point presently.

After the main capacitor and the power supply, attention must be turned to the triggering conditions. In the circuit shown in fig. 2 the voltage divider is usually designed with $R_2 \leq R_1$, so that where the operating voltage is 500 V we must reckon with a primary triggering voltage of less than 250 V (this is the voltage on the contact in the camera).

The decisive factor here is the available triggering energy. To guarantee a reliable discharge, the designer of electronic-flash lamps requires a triggering energy of at least 2 mWsec. For reasons of safety an upper limit of 12 mWsec has been proposed (see the recommendation in reference [6]). The triggering energy $W_\mathrm{i} = \frac{1}{2}C_\mathrm{i}U_\mathrm{i}^2$ is established by the suitable choice of the primary triggering voltage U_i and the capacitance C_i of the triggering capacitor. Present-day flash units remain far below the safety limit, and generally deliver a triggering energy between 3 and 5 mWsec, but not more, in order to avoid loading the contact in the camera more than necessary. The loading of the contact has been investigated, and a value recommended for the product of maximum current and maximum voltage on the contact which should not be exceeded if the contact is to have a long life. This condition leads to the specification of a minimum value for the inductance L_i of the trigger transformer. From $\frac{1}{2}L_\mathrm{i}i_\mathrm{max}^2 = \frac{1}{2}C_\mathrm{i}U_\mathrm{i}^2$ the value of the maximum current is $i_\mathrm{max} = U_\mathrm{i}\sqrt{C_\mathrm{i}/L_\mathrm{i}}$, giving the condition $U_\mathrm{i}^2\sqrt{C_\mathrm{i}/L_\mathrm{i}} \leq 2250$ W. In designing the trigger units, use is commonly made of a nomogram such as that in *fig. 3*, from which the value of C_i and the minimum value of L_i can be read off for a given U_i and W_i.

The triggering energy is converted into a very short pulse, which covers approximately half a cycle of the sinusoidal oscillation produced by the circuit consisting of C_i and the inductance L_i in the triggering unit. The length of the pulse is thus $T_\mathrm{i} \approx \pi\sqrt{L_\mathrm{i}C_\mathrm{i}}$ sec. From the condition $U_\mathrm{i}^2\sqrt{C_\mathrm{i}/L_\mathrm{i}} \leq 2250$ W it then follows that $T_\mathrm{i} \leq \pi W_\mathrm{i}/1125$ sec, so that with a triggering energy of 2 mWsec the available pulse length should be at least 6 μsec, and with $W_\mathrm{i} =$ 12 mWsec at least 35 μsec. To get some idea of the triggering time of a flash lamp, it must be borne in mind that strong ionization occurs only when the voltage of the pulse is near maximum, and therefore the effective ionization time may only be half or one third of the total pulse length.

The flash-lamp designer requires from the triggering circuit not only a minimum energy W_i but also a certain minimum triggering potential, i.e. a minimum secondary voltage at no-load operation of the transformer. The value generally required is 8 kV, which ensures reliable triggering. Oscillographic investigation of the triggering process has shown that the secondary voltage does not usually reach this value in practice, because the triggered flash lamp presents a short-circuit path to the triggering voltage. The oscillograms also show that, after the actual triggering process is completed and the discharge in the lamp has taken place and extinguished, a number of damped oscillations still occur in the triggering circuit; since the main capacitor is discharged, however, these can never cause the lamp to fire again.

A trigger unit based on the circuit of fig. 2 has been designed by Philips [7]), which meets both the energy and voltage requirements mentioned above and is very simple in design (printed wiring is used). It is fitted with an extremely reliable transformer, having polystyrol insulation and a ferroxcube core, see *fig. 4*.

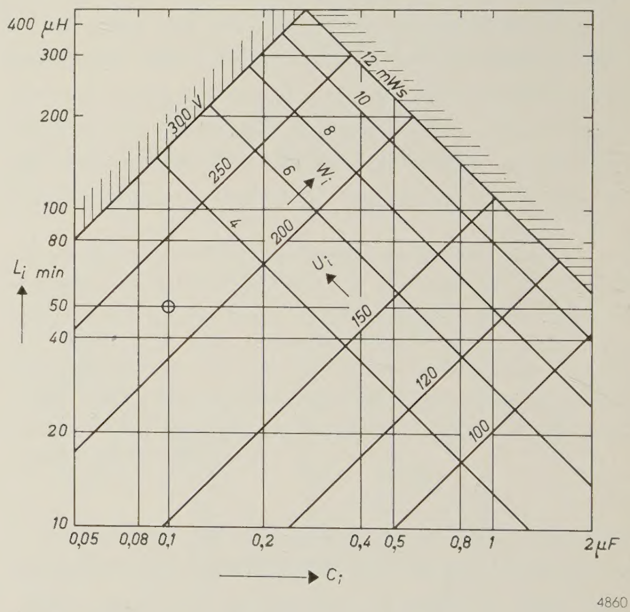

Fig. 3. Nomogram for the design of the triggering circuit C_i-T in fig. 2. At a given triggering voltage U_i and energy W_i, the diagram gives the required capacitance C_i for the triggering capacitor and the minimum inductance $L_\mathrm{i\,min}$ required for the transformer. The point corresponding to the triggering unit shown in fig. 4 is marked with a ring: $C_\mathrm{i} = 0.1$ μF, $L_\mathrm{i} = 50$ μH.

7) Developed by H. E. van Brück and C. Slofstra of Philips Icoma Division (Industrial Components and Materials), Eindhoven.

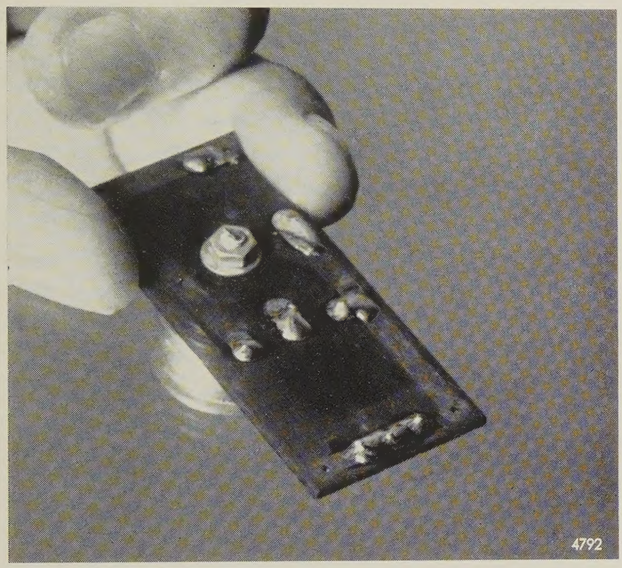

Fig. 4. The Philips triggering unit, with printed wiring. The trigger-pulse transformer has a coil wound on polystyrol round a ferroxcube core.

We shall now briefly consider the relation between the operating voltage and the triggering voltage of an electronic-flash lamp. As the operating voltage is raised, the necessary triggering voltage — which of course is always higher than the operating voltage — decreases. This is represented graphically in *fig. 5*, except that instead of the triggering voltage, which appears on the secondary of the transformer, the primary triggering voltage is shown (this is proportional to the triggering voltage). The regions "ignition" and "non-ignition" are not divided by a sharp line, but by a transitional region of some breadth. In any given flash lamp there is always a certain spread in the triggering process, and a still greater spread between different lamps of the same type. The diagram in fig. 5 was obtained by varying the operating voltage and the triggering voltage independently of one another (in most flash units the available primary triggering voltage is proportional to the operating voltage). Fig. 5 shows that

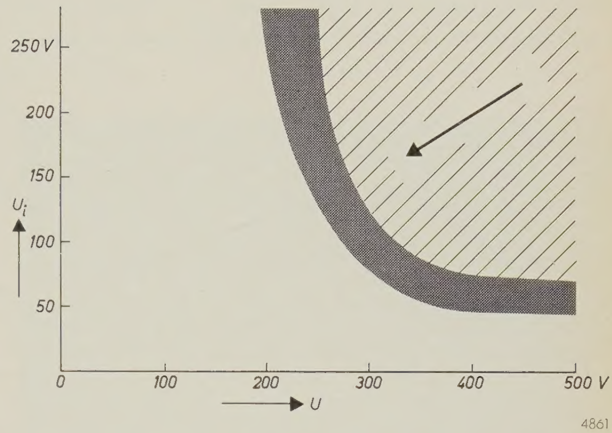

Fig. 5. The required primary triggering voltage U_i of an electronic-flash lamp as a function of the lamp operating voltage U. At higher lamp voltages, lower triggering voltages are needed. The regions "ignition" (hatched) and "non-ignition" are not separated by any sharp dividing line. In a given circuit the lamp voltage and the primary triggering voltage are proportional to one another; and if they decrease, the operating point shifts in the direction of the arrow.

as the operating voltage drops — thereby shifting the operating point roughly in the direction of the arrow — the region is gradually reached where the flash lamp will no longer ignite with certainty, or will not ignite at all. To avoid this it is increasingly the practise to stabilize the primary triggering voltage with a neon lamp, or, as in some modern control circuits, to stabilize the operating voltage itself.

Spectral distribution, integrated light intensity and flash duration

The user of an electronic-flash lamp needs to know the following concerning his flash outfit: the spectral distribution of the radiated light, the light output, and the duration of the flash. Also of importance is the spatial distribution of the radiation achieved with the aid of a reflector, but we shall not be concerned with that here.

The spectral distribution of the radiated light depends in the first place on the type of gas with which the lamp is filled. The inert gas xenon, in a discharge of the kind produced in electronic-flash lamps, delivers a continuous spectrum which largely corresponds to that of "natural" daylight. This is one of the reasons why nearly all electronic-flash

lamps today are filled with xenon. *Fig. 6* shows the measured spectral distribution of the radiation emitted by a xenon-filled flash lamp, compared with the spectral distribution of daylight [8]).

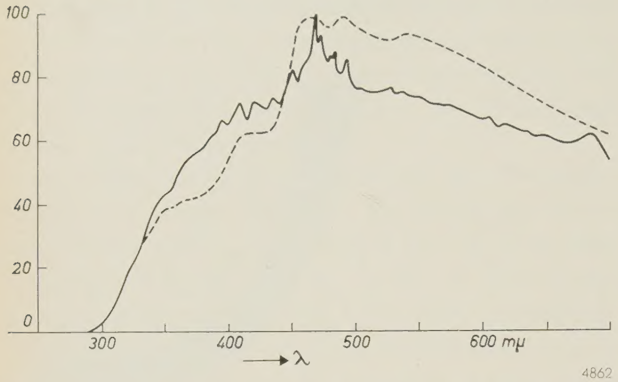

Fig. 6. Spectral distribution of the radiation from a xenon-filled flash lamp (solid line). For comparison, the broken line indicates the distribution of daylight [8]).

The fact that the light emitted by xenon-filled flash lamps so closely resembles daylight can be understood by considering the colour-temperatures of the radiations (i.e. both radiations compared with that of a black body). The colour-temperature attributable to natural daylight, which is composed of sunlight and the diffuse radiation in the firmament [8]), is roughly 6000 °K, and that found in the radiation from electronic-flash lamps is between 5800 °K and 7100 °K. The colour-temperature of flash lamps increases for a given type of gas with the specific loading of the tube wall, that is the electrical load divided by the total area of the discharge-tube walls. Xenon discharges give the best correspondence to daylight, with colour-temperatures of about 6000 °K at specific loadings in the region of 15 Wsec/cm^2, and of about 7000 °K at 30 Wsec/cm^2.

Because of the high colour-temperature of xenon-filled electronic-flash lamps, it is possible with this light to use normal daylight colour films, unlike the situation with continuously burning photographic lamps, where colour film specially sensitized for artificial (tungsten) light has to be used. The flash unit can also be used in daylight to provide supplementary lighting in dark shadows without causing impermissible colour distortion.

With regard to light output and luminous efficiency, i.e. the quantity of light per watt.second, we shall confine ourselves to the flash lamp itself without taking any account of the influence of the reflector, which is of course essential to the photographically effective use of the light.

In specifying the light output of a photographic light-source it must not be forgotten that the concept "light" by definition relates only to radiation of wavelengths between 380 and 780 mμ, and that every radiation contribution in this spectral region is evaluated in accordance with the spectral sensitivity of the eye. The units commonly used in lighting engineering are therefore based on this spectral sensitivity, for which a certain average curve has been internationally accepted. The photographic use of the light is not concerned with the human eye but with photographic emulsion. Since the spectral sensitivities may differ widely from one emulsion to another — consider, for example, orthochromatic and panchromatic emulsions, or the differing colour sensitivities of negative film and reversal film — any measurement of the "light output" would have to be based on an emulsion of average spectral sensitivity. At one time this was in fact done. Owing to various difficulties, however, including the reproducibility of such a standard emulsion, use is nowadays made of the above-mentioned system of lighting units for evaluating the light output and luminous efficiency of photographic light-sources; the light output, then, is given in lumen.seconds and the luminous efficiency in lumens per watt (or, in our case, in lumen.seconds per watt.second).

Suitable gas fillings, apart from xenon, are the inert gases argon and krypton. Elenbaas has compared these gases in a continuous discharge [9]), and has found that a xenon filling gives the highest luminous efficiency (*fig.* 7). Luminous efficiencies from 40 to 50 lumen/watt are obtained with the xenon-filled electronic-flash lamps now being made. The luminous efficiency of a given flash lamp increases with the loading somewhat as in fig. 7 (which refers to continuous discharges). *Table I* gives such data for the Philips electronic-flash lamp OF 235 Ws (Type No. 103 740) showing the light output in lumens and the luminous efficiency at various loads. We shall return presently to the influence which the *shape* of the lamp has on the luminous efficiency.

Table I. Light output and luminous efficiency of Philips' OF 235 Ws electronic-flash lamp as a function of load. The discharge tube of this type of electronic-flash lamp is made of quartz glass.

Load (Wsec)	40	60	80	100	125	200
Light output (lm.sec)	1632	2480	3410	4400	5625	9380
Luminous efficiency (lm/W)	40.8	41.3	42.7	44.0	45.0	49.9

8) R. Herrmann, Optik **2**, 384-395, 1947.

9) W. Elenbaas, High-pressure rare-gas discharges, Philips Res. Repts. **4**, 221-231, 1949.

It may be said that the duration of the flash is the most important property of an electronic-flash lamp from the point of view of the photographer. In order to get sharp photographs of scenes with moving objects, he needs the shortest possible flash. In this respect the electronic-flash lamps meet all the requirements of normal photography: the flash duration is of the order of 1 msec, which is much shorter than that of flash bulbs of the "Photoflux" type, whose flash may last as long as 30 msec. This implies, too, that a smaller total light output is needed for an exposure with an electronic-flash lamp than with combustion-type flash bulbs, for even when the fastest shutter speeds are used, the entire flash takes place while the shutter is fully open (provided the synchronization is correct): the total light output radiated is thus usefully employed. (In the case of focal-plane shutters it is of course necessary to ensure that the very short flash occurs while the shutter blind has uncovered the whole field.)

It should be pointed out, however, that extremely short flash times may be photographically unfavourable owing to the Schwarzschild effect. At given values of the product of luminous intensity and exposure time, the blackening of a photographic emulsion in the case of very long exposure times, and with the flash lamp, very short exposure times, is less than at average exposures. Where colour-sensitive emulsions are used a similar effect occurs at very short exposures, and may already be noticeable at exposures not much shorter than 1 millisecond. From a photographic point of view, therefore, the flash duration should not be shorter than 1 msec.

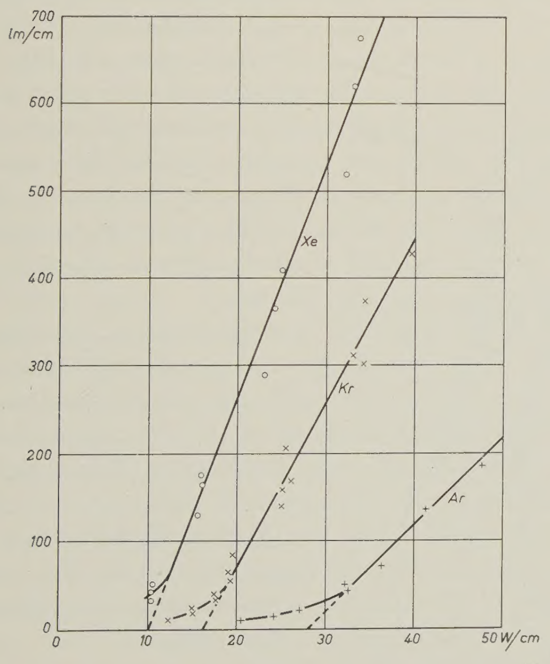

Fig. 7. Luminous flux as a function of the power in the arc of discharge tubes filled with argon, krypton and xenon, under continuous burning [9]). For the purpose of comparing discharge tubes of different dimensions, the luminous flux and power are reduced to correspond to one centimetre length of arc.

The variation of luminous flux with time can be displayed on an oscilloscope. An example of such an oscillogram is shown in *fig. 8*: a sharp rise in the luminous flux is followed by a relatively slow decline.

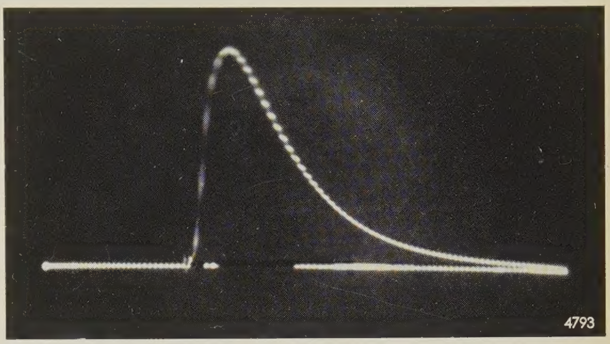

Fig. 8. Oscillogram of the luminous flux of an electronic-flash lamp. The electron beam of the oscilloscope is modulated at a frequency of 20 kc/s, so that each spot in the waveform corresponds to 0.05 millisecond.

The flash duration of various lamps, even where the time variation of the luminous flux differs from one lamp to another, may be compared on the basis of the half-width of the pulse, defined as the time during which the luminous flux is greater than half the peak value. Sometimes the 10% width is also given, the definition of which is analogous to the above. Since the triggering delay in electronic-flash lamps is negligible compared with the flash duration, it may be said that the beginning of the flash coincides with the beginning of the discharge (i.e. the triggering). In this respect electronic-flash lamps differ considerably from combustion flash bulbs.

The length of the flash depends on the design of the flash lamp, on the capacitance of the main capacitor and, as mentioned, on the operating voltage. As the capacitance increases the flash duration also gradually increases, but it drops as the operating voltage increases. The change to flash lamps with a lower operating voltage, using correspondingly larger capacitances, therefore amounted in fact to prolonging the flash duration. *Fig. 9* shows the dependence of the flash half-width on the capacitance and on the operating voltage, at a constant load of 40 Wsec. Also, for increasing load, the operating voltage remaining constant, the flash duration increases; this appears from *Table II*, which refers to a typical electronic-flash lamp. It can be seen that at the nomi-

nal value of the operating voltage the flash durations (half-widths) are still shorter than a millisecond. They are thus shorter than are really desirable from a photographic standpoint. This is a consequence of the geometrical limitations imposed on the flash

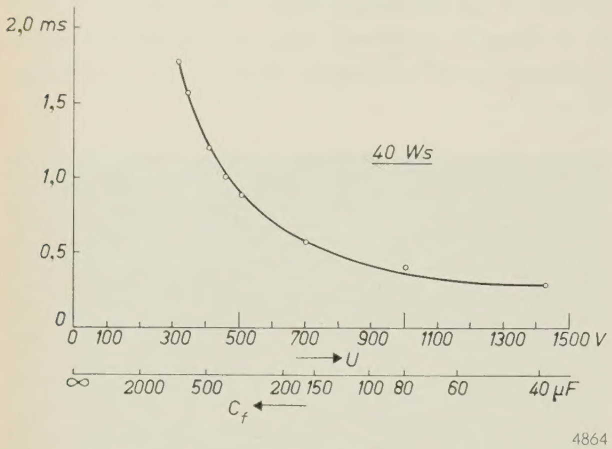

Fig. 9. Relation between the half-width of the pulse and the capacitance C_f and the operating voltage U, under constant load. Longer flash times are obtained the lower the voltages used and the higher the capacitances.

lamps: considerations of simple optics set a limit to the arc length and to the diameter of the discharge tube. At the present time it is not yet technically possible to achieve half-widths of 1 millisecond or more with the desired convenient dimensions of flash unit and reflector.

For some purposes, for example for optical warning signals on police cars, a particularly long flash is required. This is possible by sacrificing something of the luminous efficiency, the compactness and the cheapness of the flash apparatus; one can, for example, connect a resistance in series with the flash lamp, or modify the circuit in accordance with the considerations given earlier.

Table II. Flash duration (measured as half-width and 10% width) of a typical electronic-flash lamp as a function of load, at a constant operating voltage of 485 V.

Flash energy (Wsec)	60	125	200
Half-width (msec)	0.50	0.67	0.84
10% width (msec)	1.11	1.62	2.29

Construction and manufacture of electronic-flash lamps

In broad lines the construction of electronic-flash lamps may be described as follows. Sealed into the glass tube are two electrodes between which the discharge takes place. The tube is filled with xenon (usually) to a pressure of a few hundred torr (1 torr = 1 mm Hg). The trigger electrode is mounted on the outside of the tube. Pins are usually fitted, serving the dual purpose of electrical connections and mounting legs. The pins may be fixed in one or more bases. These details of the construction will be further discussed presently.

The discharge tube proper is usually of hard glass, sometimes quartz glass. The latter is used where the tube wall is to be subjected to particularly heavy loading. If a great deal of heat is generated during the discharges, the glass may be locally heated to its softening point, giving rise to stresses upon cooling; in the course of time hair-cracks may form ("sintering"), which may finally lead to breakage. There is much less danger of this happening with quartz glass, owing to its higher melting point and lower expansion coefficient.

Although formerly hard glass was usually found to be adequate in most cases, increasing use has recently been made of quartz glass, in view of the growing trend towards small, heavily loaded lamps. Since a point source is optically preferable, this trend is understandable. However, as can be seen in *fig. 10*, the discharge tube in practice takes the most various forms. The simplest shape for round reflectors, which were initially very widely used, is the U shape (*a*). Somewhat more complicated, but better adapted to the round reflector, is the Ω form (*b*) or the helical loop (*c*). Efforts were later made to adapt the reflector to the rectangular form of the picture, with the idea of ensuring that the picture would receive uniform overall illumination. For this purpose a linear shape of tube is more suitable: (*d*) or (*e*), the latter with the electrodes fitted perpendicular to the tube. Finally there is the spiral or helix form (*f*), which has been popular for high-tension apparatus right from the beginning. Since a long discharge path was needed for high operating voltages, the obvious method of producing a compact light source was to spiralize the tube.

In conjunction with a suitable reflector, the shape of the lamp can influence the spatial distribution of the light. The luminous efficiency, on the other hand, depends on the dimensions of the tube (apart from the electrical operating conditions). For a given flash energy, the dimensions of importance are the inside diameter of the tube and the distance between the electrodes, i.e. the length of the tube. We shall return to this question when we come to consider the optimum design of a flash lamp.

Turning now to the electrodes, which are sealed into the ends of the discharge tube, there are two processes that take place at the electrodes and therefore determine their design. The first is the triggering process (ignition), the second is the actual burning

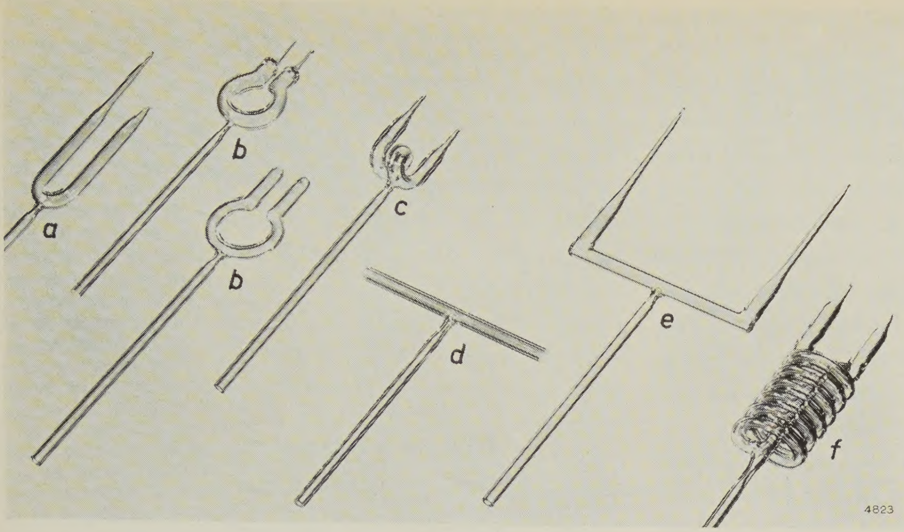

Fig. 10. Various shapes of discharge tube.

of the lamp which, though short-lived, can cause considerable heating of the electrodes as a result of the high discharge current. The electrodes must be designed in such a way as to enable the tube to ignite at given values of operating and triggering voltages, and they must continue to do so during the whole life of the lamp. Furthermore, the evaporation of electrode material should be minimized to prevent deposits forming on the wall and reducing the light output during the life of the lamp.

In order not to be unduly restricted in the choice of the other parameters, a low triggering voltage is aimed at, which implies that the electrode material must have a low work function. To this end the electrodes, which consist of a tungsten wire core with a tungsten or molybdenum wire wound around it, are coated with a highly emissive substance. In most cases the emitter substance consists largely of thorium oxide and barium oxide. The emitter also acts as a getter. It has been found that the anode, too, can advantageously be given an emissive coating. The two electrodes are therefore usually made identical. It is not advisable, however, to operate flash lamps using one electrode first as cathode and later as anode, or vice versa. The electrodes must therefore be distinguishable. In the type of flash lamp described here the electrode used as cathode during the burning-in and aging periods in manufacture can be recognized from the fact that the connection to the trigger electrode is situated at the cathode end of the flash lamp (see e.g. figs. 11 and 15).

The kind of trigger electrode used also has its influence on the ignition of the lamp. In electronic-flash lamps the triggering is invariably capacitative, i.e. the trigger electrode is mounted on the *outside* of the discharge tube. As can be seen in *fig. 11*, there are three common types of trigger electrode: a spirally wound wire; a conducting strip usually laid parallel to the discharge along the inside of the loop; and a transparent, conductive layer with which the entire discharge tube is coated. The discharge is initiated along the trigger electrode, and for this reason the conductive layer, introduced in recent years, has the advantage over the others.

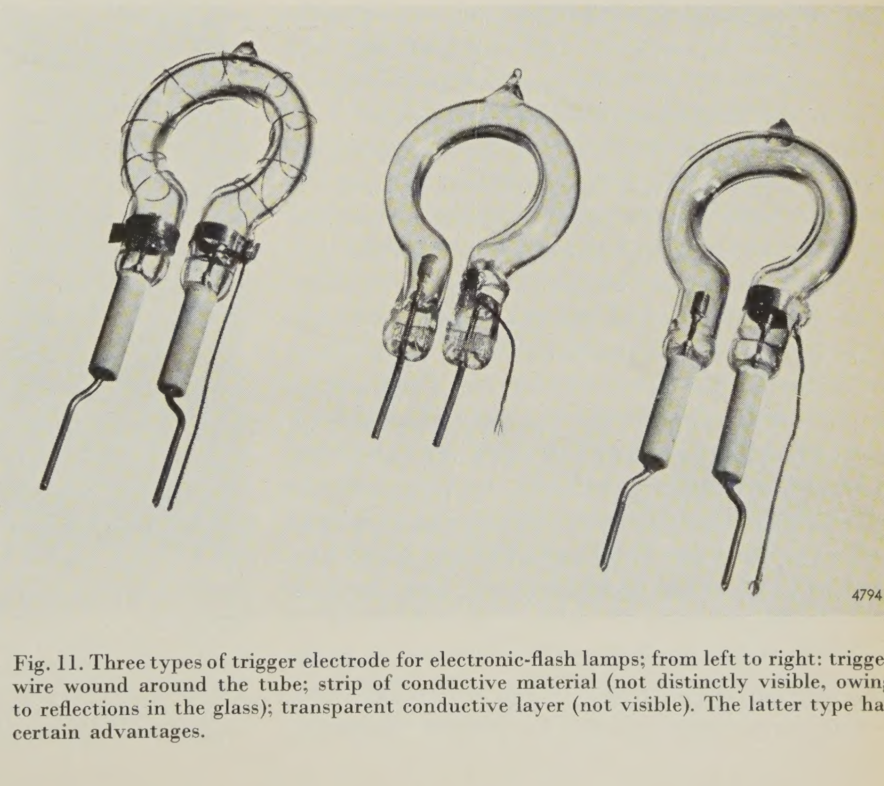

Fig. 11. Three types of trigger electrode for electronic-flash lamps; from left to right: trigger wire wound around the tube; strip of conductive material (not distinctly visible, owing to reflections in the glass); transparent conductive layer (not visible). The latter type has certain advantages.

It gives the lowest triggering voltage and leads to a uniform discharge throughout the tube.

A word here about the sealing-in of the electrodes. Generally speaking the seals present no difficulties, in spite of the high current surges of several hundred amperes. In quartz-glass tubes, however, where the usual pinch seal with thin molybdenum foil is to be used, the current surges can cause considerable heating at the seals and may destroy the thin foil used. This problem can be solved by using somewhat thicker molybdenum foil. In some particularly troublesome cases, tungsten seals with intermediate glass had to be adopted [10]).

The economic manufacture of electronic-flash lamps is somewhat of a problem owing to the relatively small production runs required. For this reason, flash lamps are frequently hand-made. Nevertheless, even in the case of small production runs, some mechanization is an advantage in that it leads to a more uniform quality of the product. To what extent is mechanization possible even in the manufacture of small series of flash lamps? In many cases the discharge tube can be bent to the required shape mechanically. A simple machine for this purpose is shown in *fig. 12*. The glass tube is heated by a burner (*A*), rough-shaped around a mandrel (which projects from the bottom of pipe *B*) and introduced into a mould (*C*), after which the two halves (*D*) of the mould are closed and air is blown into the tube (the other end of the tube being sealed off). This method of shaping has the advantage of keeping the outer dimensions of all tubes within very narrow limits, a point of importance as regards assembly in the reflector. After a pump stem has been sealed to it, the discharge tube is thoroughly cleaned and dried ready for further working.

Parallel with these operations, the electrodes are made. To begin with, the electrode wire core is provided with a bead of the intermediate glass (*fig. 13a*), which serves to compensate for the difference in expansion coefficient between the electrode wire and the glass tube (we shall discuss here only the case of hard-glass tubes). This being done, the helix of W or Mo wire is slid onto one end of the glazed core, and the complete assembly is again thoroughly cleaned and dried (fig. 13*b*). The electrodes are then coated with emitter paste, in such a way that the paste also penetrates into the space between the core and the helix wire (fig. 13*c*). After the electrodes have been sintered in a tungsten-strip furnace, their surface is carefully brushed (fig. 13*d*). The discharge should not spring directly from the emitter substance, since this might cause evaporation of the latter and lead to the formation of light-absorbent deposits on the glass wall. On the other hand, the complete absence of emitter material on the outside of the electrode would imply a higher triggering voltage. A compromise is therefore adopted, some emitter material being left

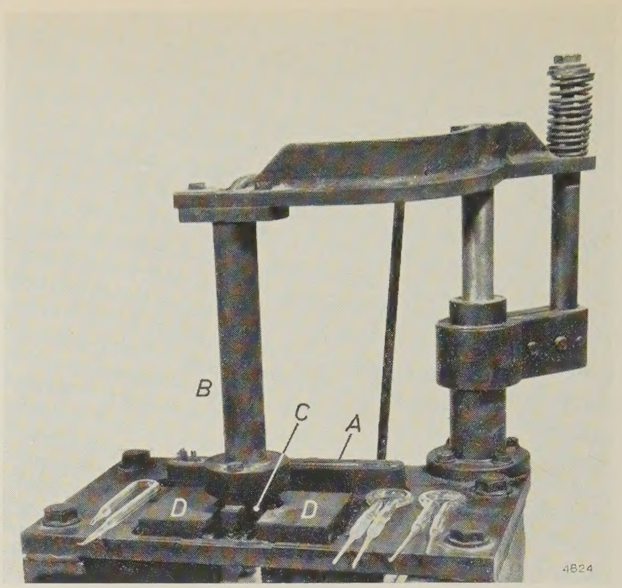

Fig. 12. Simple machine for shaping the envelopes of discharge tubes, in this case Ω-shaped. *A* burner for heating the glass tube. *B* pipe, through the base of which a mandrel projects for pre-shaping the glass tube. *C* mould into which the tube, bent into a U shape, is introduced. *D* mould halves for final shaping.

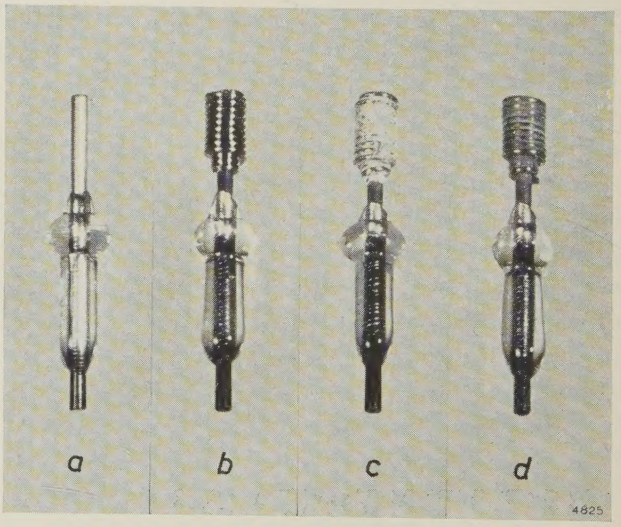

Fig. 13. Stages in the fabrication of electrodes for hard-glass flash lamps.
a) Wire core with intermediate glass.
b) Wire helix fitted over one end of the core.
c) Assembly coated with emissive paste.
d) After sintering in a tungsten-strip furnace, the superfluous paste is brushed away from the spiral.

[10]) See in this connection page 81 of the article by P. Hoekstra and C. Meyer, Motion-picture projection with a pulsed light source, Philips tech. Rev. **21**, 73-82, 1959/60 (No. 3).

on the outside of the electrodes, which vaporizes when the lamp is burnt in and forms a slight deposit on the glass wall — a deposit which, however, causes no significant drop in light output.

The two electrodes can now be sealed into the discharge tube. This too is done on a small machine (*fig. 14*), making it possible — since both electrodes are sealed in at the same time — to achieve the specified dimensions, in particular the electrode spacing, more accurately and faster than by hand. Next, the tubes are evacuated on the pump and carefully degassed. Gases or vapours released from the electrodes or glass wall may contaminate the gas filling and make the tube unreliable in operation. With this in mind, the tube is evacuated to a pressure lower than 10^{-4} torr. If necessary, the electrodes can then be activated, and the tube is filled with xenon to the specified pressure and sealed off. The subsequent operations consist primarily in applying the trigger electrode. If this is a conductive

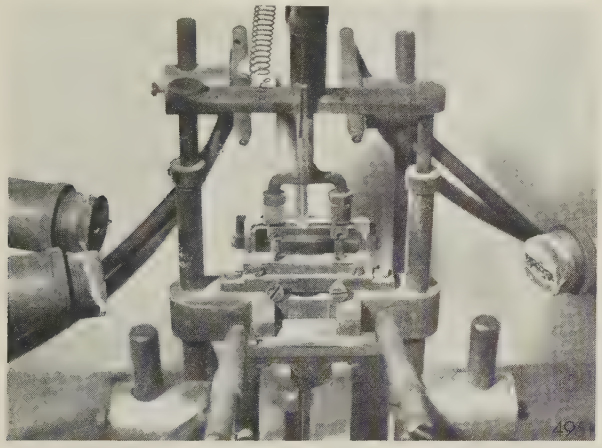

Fig. 14. Machine on which the two electrodes of an electronic-flash lamp are simultaneously sealed into the tube.

layer, it is sprayed on in an oven whose temperature is nearly at the softening point of the glass. The connection wires are then soldered to the tube and a base is fitted if required. This completes the actual production process (*fig. 15*).

The flash lamps have still to be aged and inspected, however. Both steps are of especial importance for obtaining a product of good quality. Aging is necessary to bring the electrodes into the best condition for operation. We have seen that the emitter is not only important as regards reliable ignition but also acts as a getter. Although it is not possible to say exactly which part of the electrode is responsible for ignition and which part for gettering, it has been found that, after a certain number of flashes under normal load, the flash lamp becomes stable in opera-

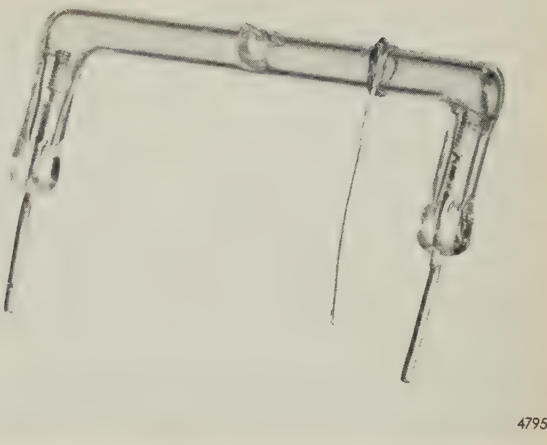

Fig. 15. Philips electronic-flash lamp, type OF 50 Ws (No. 3 in fig. 1).

tion, which means in particular that the triggering voltage becomes fairly constant. In the inspection stage the principal dimensions of the lamps are examined, the triggering voltage of each lamp separately is measured, and sample inspections are made to determine the light output, and also the decline in light output after a specific number of flashes.

The optimum design of electronic-flash lamps

So far various questions that arise in the development of electronic-flash lamps have been examined separately. We shall now consider the question of how to produce "optimum" flash lamps, and to do this we must investigate the relation between the electrical and lighting requirements and the constructional details.

The three essential requirements to be met by a flash lamp have already been mentioned: reliable triggering, maximum light output under given conditions, and minimum decline in light output during the life of the lamp. The electrical circuit to be used is generally established. As regards triggering, then, we may assume that the following data are given: the minimum operating voltage at which the lamp must ignite, the minimum primary triggering voltage available, the primary triggering energy available, and the type of trigger-pulse transformer to be used. These data are needed, and the triggering transformer itself must be available, in order to determine the ignition behaviour of a flash lamp. As regards the voltage data, it must also be borne in mind that a certain safety margin is necessary in the production process. In ascertaining whether the specification has been met, we must therefore take values a few per cent lower than the minimum values mentioned.

Calculation of luminous efficiencies from measurements of the light output must be done on the basis of the nominal values of the operating voltage and the capacitance of the main capacitor.

Life tests, for determining the number of flashes that the lamp can withstand, must be done at the maximum load likely to be encountered in practice, that is to say the highest encountered values of capacitance and operating voltage. The decline in light output during the life of the lamp is also determined under these conditions, and used as the basis for assessing the quality of the lamp. On the other hand, the light-output measurement necessary for determining this decline is carried out under nominal loading.

It is also necessary to know the maximum operating voltage because of the fact that, if the voltage used is too high (close to the self-breakdown voltage), the flash lamp may ignite without being triggered. It is particularly necessary to take this into account in the case of high-tension flash lamps, where the voltage of the high-tension supply may vary considerably.

We have seen that marked differences between actual and nominal voltages can cause a considerable spread in the light output of the lamp. The same applies to differences in capacitance and resistance values in individual flash units of the same type. Capacitors, for example, can normally only be made with a tolerance of -10% to $+20\%$ in their capacitance value. This again entails a spread in the light output by more than 25%. Because of this capacitance spread and the differences already mentioned in the charging potential of the capacitor, the user of a flash unit has hitherto had to make a series of test exposures to determine the right stop to be selected or the guide number to be used. In this respect the advent of voltage-stabilized circuits has brought some improvement, but there is still the need to have all electrical data accurately specified without unduly wide tolerances. As regards the capacitance tolerances of electrolytic capacitors there has latterly been some slight improvement.

Once the electrical data are established, the behaviour of the lamp in respect of triggering and light output can be controlled by varying the dimensions of the discharge tube, i.e. the electrode spacing l and the inside diameter d, and also by varying the pressure p of the gas filling.

The relation between these quantities has been experimentally investigated on linear discharge tubes; inclusion of the effects of tube shape would have endlessly prolonged the investigation. In any case, the data for bent tubes are not reliable, inasmuch as bending changes the diameter of the tube in a manner that is difficult to define satisfactorily. The experiments were set up according to statistical principles. The various parameters were varied within the following limits: electrode spacing from 20 to 80 mm, diameter of tube from 2 to 6 mm and gas pressure from 100 to 700 torr. Preliminary measurements had shown that the maximum light output was to be expected between these limits. The entire investigation was done in three experimental runs, namely for three different loads of 40, 62.5 and 100 Wsec. Although of course the light output from an electronic-flash lamp increases with increasing load, the load itself (i.e. the flash energy) is laid down by the flash-unit designer and is therefore not a freely variable parameter.

The light output, with constant tube diameter and constant load, is found to depend on the electrode spacing l and the gas pressure p in the manner shown in *fig. 16*. We see that there is indeed a cer-

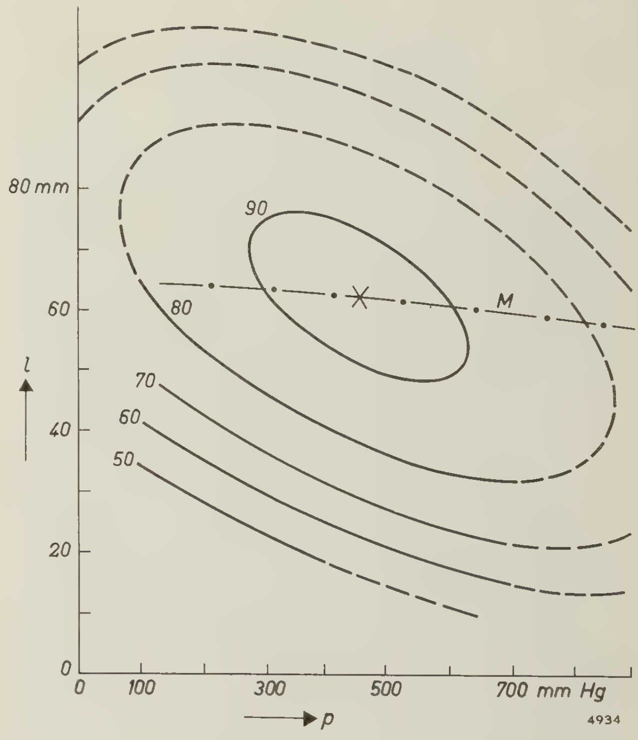

Fig. 16. Curves of constant light output of a flash lamp showing dependence on the gas pressure p and the electrode spacing l, for constant discharge-tube diameter d and constant load. The figure on each curve represents the light output expressed as a percentage of the maximum. (The dashed portions of the contours were obtained by extrapolation [11].) If the diameter d of the discharge tube is varied, the point of maximum light output shifts along the chain line M; at a certain point along this line (i.e. at a certain tube diameter) there is an optimum maximum light output, i.e. a maximum of the various maxima.

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[11]) Statistical analysis of the results carried out by J. W. Sieben (Lighting Division, Eindhoven).

tain optimum combination of p and l at which the light output is a maximum. When the tube diameter is varied, the position of the peak shifts along the chain line M, and the height of the peak itself becomes maximum at a particular point along this line. Of course, when the tube diameter is varied, the whole family of curves shifts together with the position of the peak. Since this involves no fundamentally new phenomena, however, it is not necessary to reproduce all these diagrams here. The same applies to the diagrams for the three different loads.

It follows from the measurements just described that it is indeed possible, with given electrical conditions, to produce an optimum flash lamp, provided the tube diameter, the electrode spacing and the gas pressure may be freely chosen. If some other shape of flash lamp is required, e.g. a U shape or Ω shape — we assume that the results obtained on linear discharge tubes can be applied to other shapes without serious error — the lamp designer can make one optimum type of flash lamp for any required load. The dimensions may be adduced from fig. 16 or from the corresponding diagrams for other tube diameters and loads.

It may happen that a flash lamp thus designed is found to be too large and that for optical reasons, i.e. with an eye to the reflector, a smaller and more compact lamp is required (in most cases shorter). Furthermore, a small tube diameter may be wanted in order to lengthen the duration of the flash. In that case, only the gas pressure can be freely chosen, and here too the best choice may be adduced from a diagram as in fig. 16.

We still have to ascertain the way in which triggering is affected by the variation of parameters l, d and p. As a rule, the minimum operating voltage at which the flash lamp can still just be ignited increases with rising l and p and with decreasing d. This relation can be represented by contours of constant minimum triggering voltage in an l-p diagram as shown in *fig. 17* for the same case as in fig. 16 (i.e. for the same tube diameter d). The figure shows, for example, that maximum light output cannot be achieved with a flash-unit circuit where the lower limit of the available primary triggering voltage is 260 V: to ensure reliable ignition in that case, we must make l and/or p smaller than the values needed to produce maximum light output.

We have not yet considered the decline in light output during the life of the lamp. It is mainly governed by the quality of the electrodes, and is not much affected by the three parameters l, d and p.

Finally, an idea of the present situation in the development of electronic-flash lamps is given in *Table III*, which gives data for a few representative types. A few years ago, most flash units for amateurs were still equipped with lamps for a flash energy of about 80 to 120 Wsec. Nowadays a light output sufficient for most photographic purposes is obtained with a flash energy of 45 Wsec. Examples of such flash lamps are type OF 45 Ws (linear shape) and OF 50 Ws (fig. 15; electrodes perpendicular to discharge tube to reduce length). The U-shaped lamp OF 165 Ws is intended for higher-performance apparatus, and can be subjected to a maximum load of 165 Wsec. The light output from this lamp is so high that objects larger than normal can be given good overall illumination; for close-up shots of smaller objects it is then sometimes necessary to change to a lower flash energy.

The two types OF 80 Ws and OF 100 Ws approach closely to the optimum design described above, at loads of 80 and 100 Wsec, respectively. Type OF 235 Ws, with Ω-shaped quartz-glass discharge tube, is intended for press photographers. Type OF 500 Ws is a universal flash lamp, of which only

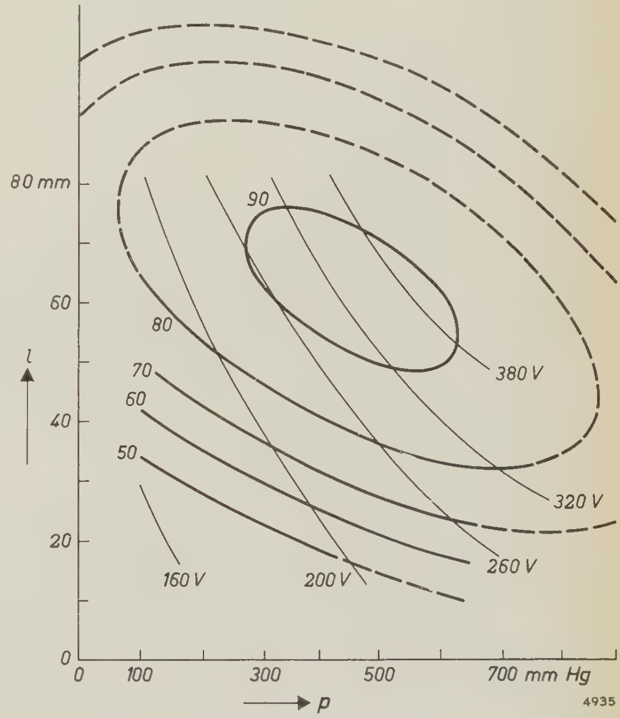


Fig. 17. At a given tube diameter the minimum triggering voltage can be specified for any electrode spacing l and gas pressure p. The set of thin curves shown here is found by joining together the points of minimum triggering voltage in fig. 16. If the power-supply circuit in a flash unit is so designed that the (minimum) available triggering voltage is only 260 V, the maximum light output cannot be achieved in the case shown here: electrode spacing and gas pressure would then have to be chosen small enough to remain on the left of the curve for 260 V.

Table III. Data on some Philips electronic-flash lamps: in each case the type designation indicates the maximum permissible flash energy.

Type designation and number	Operating voltage V	Shape	Envelope	Electrode spacing mm	Light output lm.sec	Luminous efficiency lm/W	Wall load per flash Wsec/cm²	Colour-temperature °K
OF 45 Ws 103 909	500	fig. 1, No. 1	glass	44	1900	47.6	13	6400
OF 50 Ws 103 931	500	fig. 15						
OF 165 Ws 103 798	500	fig. 1, No. 9	glass	70	6550	46.8	12	6500
OF 80 Ws 103 952	500	fig. 1, No. 12	glass	65	2735	45.6	7.3	5800
OF 100 Ws 103 958	500	fig. 1, No. 10			2830	47.2		
OF 235 Ws 103 740	500	fig. 11	quartz glass	63	9380	46.9	23	7100
OF 500 Ws 103 965	500 to 2500	fig. 1, No. 5	quartz glass	110	15 000	50	22	6600
OF 1100 Ws 103 752	2700	fig. 10f	quartz glass	500	39 500	49.4	18.5	5800
OF 1500 Ws 103 730	2700	fig. 1, No. 4	quartz glass	205	45 600	45.6	38.8	6900

experimental versions have been made, and which operates reliably in a very wide range of voltages: the operating voltage can be selected in this case between 500 and roughly 2500 V. Because of the relatively large dimensions of the discharge tube, this type provides the relatively long flash duration required for photography: at lower voltages half-widths of a few milliseconds are obtained.

For comparison, the table also gives data on two representative types of high-voltage electronic-flash lamps. The spiralized lamp OF 1100 Ws can be used in flash apparatus which, though heavy, is nevertheless portable. Type OF 1500 Ws is intended for studio use; shaped like a large ring, it can be fitted around the lens of the camera.

Summarizing, the major development in the design of electronic-flash lamps may be said to have been the reduction of the flash energy for amateur equipment to 45 Wsec, made possible by the improvement of luminous efficiency. The introduction of these 45 Wsec lamps has made it a practical proposition to produce electronic-flash units weighing less than 2 pounds. The general trend is in the direction of still lower flash energies with a view to producing even smaller and lighter flash equipment. Present indications, however, are that the luminous efficiency of the lamps will then be lower.

Summary. Electronic-flash lamps are made in a wide variety of types to meet the requirements of the manufacturers of flash equipment. An important feature of developments in this field in recent years has been the advent of flash lamps for operation at the relatively low voltage of 500 V. This has made it possible to produce readily portable flash equipment fitted with light-weight electrolytic capacitors. The designer of flash lamps must work on the basis of both the electrical operating conditions (governed by the triggering and discharge circuit) and the photographic lighting requirements (spectral distribution of light, light output and flash duration). Lamps filled with xenon (at a pressure of some hundreds of torrs) give a spectral distribution closely resembling natural daylight. The light output at a given flash energy (luminous efficiency) has been so improved in recent years that a flash energy of 45 Wsec is now sufficient for most photographic purposes. As a result it is now for the first time possible to make flash apparatus with a total weight of less than 2 pounds. The duration of the flash is short enough for most practical cases; indeed, in view of the Schwarzschild effect, it was even desirable to make it somewhat longer, and the flash duration has now been brought close to the desired value of 1 millisecond. After a description of the basic design and methods of manufacture of flash lamps, details are given concerning the appropriate choice of the electrode spacing, the diameter of the discharge tube and the gas pressure to produce an optimum design. Tabulated data are given for a few typical electronic-flash lamps made by Philips.

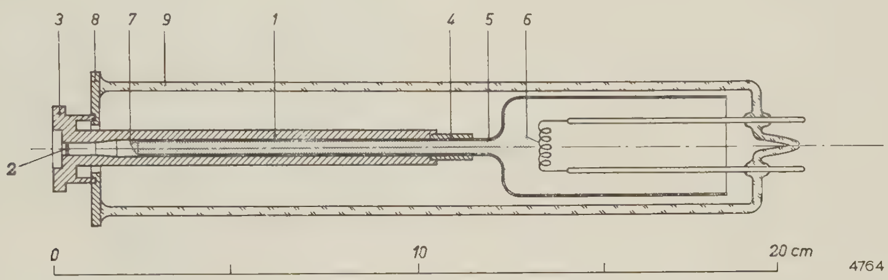
AN EXPERIMENTAL NOISE GENERATOR FOR MILLIMETRE WAVES

621.373.432.029.65:621.327.52

A microwave noise source is needed for purposes such as measuring the noise factor of millimetre-wave equipment (e.g. radar receivers), and also as a standard noise source in plasma research. A suitable noise source of this nature is the positive column of a discharge in an inert gas, provided it is so dimensioned as to give a high equivalent noise temperature. External factors, like the magnitude of the discharge current, the filament voltage and the ambient temperature, generally have little influence on the noise temperature and the matching.

In the millimetre-wave region the positive column can be regarded as a black-body radiator of very high temperature, closely approximating to the electron temperature of the plasma. The electron temperature is primarily governed by the gas used — at least under the conditions chosen for the discharge column in a noise generator — and is higher the lighter the gas atoms.

For frequencies up to about 40 Gc/s (wavelengths down to 7.5 mm) a noise generator can be made by simply taking the glass tube in which the gas discharge takes place and passing it obliquely through a waveguide in such a way that the part of the column inside the guide is properly matched and thus delivers the maximum noise power to the guide [1]). At higher frequencies, however, the dimensions of the waveguide are too small to make this system practicable. More suitable in this case is the design illustrated in *fig. 1*, which gives good results in the 4 mm band. A circular copper waveguide *1* is closed at one end (left) by a mica window *2*, and provided at that end with a flange *3* for coupling to the rest of the circuit. Inside the waveguide a thin-walled tube *5* of quartz glass is introduced. At the right this flares out into a widened section which contains an oxide cathode *6* for the gas discharge; the tube is filled with neon. The left-hand end of the tube is open and the anode is formed by the inside of the waveguide near *7*, immediately beyond the end of the tube. The positive column is contained



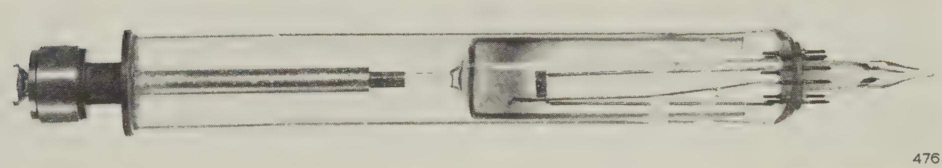


Fig. 1. Cross-section and photograph of experimental noise generator for the 4 mm waveband. *1* circular waveguide. *2* mica window. *3* flange for coupling to circuit. *4* spring clips. *5* quartz-glass tube. *6* oxide cathode. *7* part of inside wall of waveguide acting as anode. *8* molybdenum disk. *9* neon-filled glass envelope (pressure 10 cm Hg).

[1]) W. W. Mumford, A broad-band microwave noise source, Bell Syst. tech. J. **28**, 608-618, 1949.
K. S. Knol, Determination of the electron temperature in gas discharges by noise measurements, Philips Res. Repts. **6**, 288-302, 1951.

inside this tube, i.e. in the axial direction of the waveguide. The column can be made long enough for the equivalent noise temperature to approach closely to the electron temperature, so that the maximum noise power is delivered to the waveguide and is transmitted through the mica window to the rest of the circuit. To minimize reflection losses, the window is provided with a tuned diaphragm.

In principle it would be possible to seal the discharge tube at the cathode side hermetically by using a quartz-glass base with lead-ins for the cathode. Since the tube has to be very thin-walled, however, the assembly would then be too vulnerable. For this reason a design as shown in fig. 1 was adopted: soldered to the outside of the waveguide is a molybdenum disk *8*, sealed to which is a glass envelope *9* carrying the cathode leads.

The experimental tube built in this way has a neon pressure of 10 cm Hg. The discharge current is 75 mA, the burning voltage 150 V, and the noise temperature T is 21 000 °K, the maximum error of measurement being $\pm$1000 °K. The noise power available in a narrow frequency band Δf is $kT\Delta f$; k is Boltzmann's constant $= 1.38\times10^{-23}$ J/°K.

In *fig. 2* the standing-wave ratio s, measured on the experimental noise generator, is plotted as a function of frequency f. From s the percentage by which the equivalent noise temperature is lower

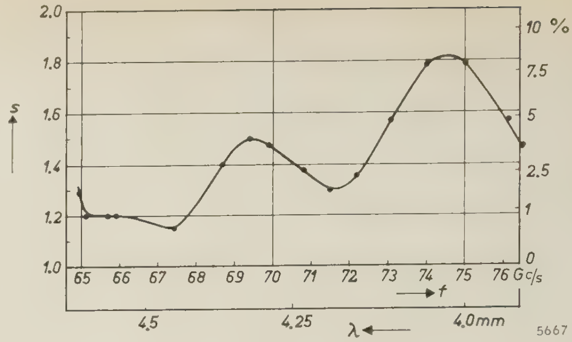


Fig. 2. Standing-wave ratio s, measured on the noise generator shown in fig. 1, as a function of frequency f (or wavelength λ). The scale on the right indicates by what percentage the equivalent noise temperature at a given frequency is lower than 21 000 °K.

than 21 000 °K at a given frequency can be derived. This percentage is shown on the scale at the right of fig. 2.

It is certain that this type of noise generator may be extended to wavelengths shorter than 4 mm. To produce a noise source for the 2 mm band it will probably be sufficient to add a transition section from 4 to 2 mm and to compensate for the mismatch.

The tube discussed here will be dealt with in more detail in an article on standard noise sources to be published in this journal.

P. A. H. HART and G. H. PLANTINGA.

MULTIPATH TRANSMISSION EFFECTS IN FM RECEPTION AND THEIR SIMULATION IN THE LABORATORY

by J. KOSTER *).

621.391.826.2:621.376.33

In mountainous regions the waves from a broadcasting transmitter may reach the receiver along multiple paths of different length as a result of reflections from mountain ridges. The consequence of this in frequency-modulated broadcasts may be severe distortion of the sound. The author has designed a signal generator for simulating and studying this interference in the laboratory. This makes it possible to check at any time, irrespective of receiving conditions, the effect of measures taken in the receiver to reduce this distortion.

In FM broadcasting (frequency-modulated VHF transmissions) use is made of waves in the metre bands. These waves as a rule reach the receivers along the direct path from transmitting to receiving aerial. Not infrequently, however, one or more other transmission paths may exist at the same time, owing to the waves being reflected from some natural obstacle, such as a mountain ridge. The various paths will generally differ in length, which means that two or more waves having different transit times and hence a phase difference arrive at the receiving aerial. The consequence, particularly if the reflected waves are not much weaker than the direct wave, is a peculiar distortion of the detected signal. The impression one receives is as if the output amplifier were overloaded, or as if something were loose in the loudspeaker. The cause, however, is of quite a different nature, as will appear from the analysis given below.

The distortion in question was very soon noticed when frequency modulation first began to be used. Its cause was also correctly ascertained, and measures for improvement were proposed [1] [2] [3] [4]. In this connection the investigations led by Arguimbau made an especially useful contribution [3]. In order to study the effect of these measures, a signal generator is needed which is capable of delivering a signal corresponding to that which an aerial receives under the conditions mentioned. A relatively simple solution of this problem is described in the present article. To make clear the requirements to be met by such a signal generator, we shall first give a simplified analysis of the effects involved.

The FM receiver

Fig. 1 shows the familiar block diagram of a normal FM broadcast receiver. The audio-frequency section (A_3-L) need not be considered here. The

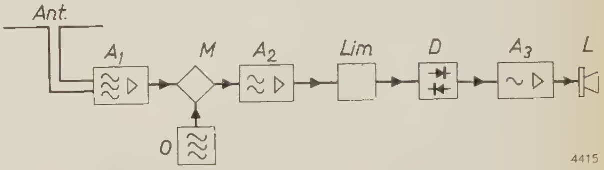


Fig. 1. Block diagram of a conventional FM broadcast receiver. *Ant* aerial. A_1 radio-frequency amplifier. M mixer. O local oscillator. A_2 intermediate-frequency amplifier. *Lim* limiter. D FM detector (discriminator). A_3 audio-frequency amplifier. L loudspeaker. (For clarity the limiter is shown as a separate block; in reality, limiting occurs partly in the last stage of A_2, partly in the discriminator.)

radio-frequency amplifier (A_1), the mixer (M) and the intermediate-frequency amplifier (A_2) can be regarded for our purposes as linear networks, in other words, we may apply to them the superposition theorem. This states that in the simultaneous presence of more than one signal the total effect is the sum of the effects of the individual signals, provided only that the amplitude characteristic is sufficiently horizontal and the phase characteristic sufficiently straight. These conditions are reasonably satisfied if the bandwidth of the amplifiers is not less than about three times the maximum frequency deviation of the transmitted signal. In the case of FM broadcasting stations the maximum frequency deviation is fixed at 75 kc/s by international agreement.

*) Radio, Television and Record-player Division, Eindhoven.

1) M. S. Corrington, Frequency-modulation distortion caused by multipath transmission, Proc. Inst. Radio Engrs. **33**, 878-889, 1945.
M. S. Corrington, Frequency modulation distortion caused by common- and adjacent-channel interference, R.C.A. Rev. **7**, 522-560, 1946.

2) F. L. H. M. Stumpers, Interference problems in frequency modulation, Philips Res. Repts. **2**, 136-160, 1947.

3) L. B. Arguimbau and J. Granlund, The possibility of transatlantic communication by means of frequency modulation, Proc. Nat. Electronics Conf., Part III, 644-653, 1947.
L. B. Arguimbau and J. Granlund, Interference in FM reception, Tech. report No. 42, Research Lab. of Electronics, Massachusetts Inst. of Technology, 1947.

4) L. W. Johnson, F.M. receiver design, Wireless World **62**, 497-503, 1956.

The limiter (*Lim*) and the discriminator (*D*) are essentially non-linear systems. It is therefore not enough to consider each input signal individually; we must also take their resultant into account. We shall see presently that in certain circumstances the instantaneous frequency of the resultant signal may make excursions far beyond the band within which the instantaneous frequencies of the constituent signals remain. To meet these unfavourable circumstances it is necessary to give the non-linear part of the receiver a bandwidth larger than is needed for a normal FM signal.

Analysis of multipath transmission effects

In an article in the previous issue of this journal it was shown that an FM receiver to which two signals are simultaneously applied will detect the stronger of the two, whilst the weaker one will act as an interfering signal [5]). Multipath reception of FM signals is to be treated as a special case of this.

To avoid unnecessary complication of the problem, we assume that the transmitter is modulated by a sinusoidal audio signal (frequency $p = \Omega/2\pi$) and that between the transmitting and the receiving aerial there are only two transmission paths, with a transit-time difference of τ. Of these two paths one may be the direct path and the other indirect, though both may also be indirect.

In the radio-frequency part of the receiver there will then be two signals present, both sinusoidally modulated in frequency and one lagging behind the other by a time τ. Instead of these RF signals we can better consider the corresponding intermediate-frequency signals, both of which are lower than their corresponding RF frequencies by the same amount (which is equal to the frequency of the local oscillator). The instantaneous frequencies f_a and f_b of the two intermediate-frequency signals are given by:

$$\left.\begin{aligned} f_a &= f_0 + \Delta f \sin \Omega t, \\ f_b &= f_0 + \Delta f \sin \Omega(t-\tau)\,. \end{aligned}\right\} \quad \ldots \quad (1)$$

Here f_0 is the centre intermediate frequency (often fixed at 10.7 Mc/s) and Δf is the frequency deviation. *Fig. 2* shows f_a and f_b as functions of time t. As can be seen, during one half of the period T of the audio signal the frequency f_a is lower than f_b, and during the other half it is higher. The difference f_d of the two instantaneous frequencies is

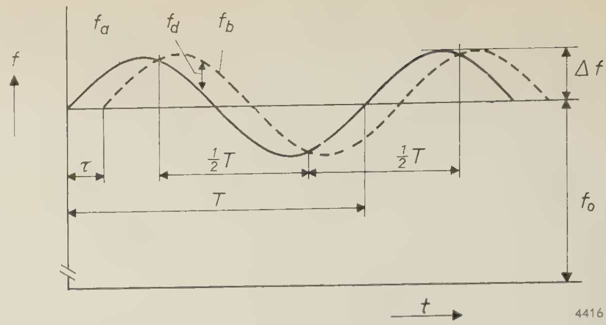


Fig. 2. Instantaneous intermediate frequencies f_a and f_b of the direct and indirect signals, respectively, received from an FM transmitter modulated by a sinusoidal audio signal (frequency $p = 1/T$). The fixed centre intermediate frequency is f_0 (usually 10.7 Mc/s). The difference in the transit times of the two transmission paths is τ.

found by simple calculation from (1) to be a cosine function of time with the audio frequency $\Omega/2\pi$:

$$f_d = f_a - f_b = 2\,\Delta f \sin \tfrac{1}{2}\Omega\tau \cos \Omega(t-\tfrac{1}{2}\tau)\,. \quad (2)$$

We represent the two IF signals by the vectors **a** and **b** (*figs. 3* and *4*) of length a and $b \leq a$, respectively [6]). We keep the vector **a** stationary; in accordance with eq. (2) the vector **b** then rotates for one half period $\frac{1}{2}T$ anticlockwise a number of times ($f_b > f_a$) and in the next half period the same number of times clockwise ($f_b < f_a$), with an instantaneous angular velocity of $\omega_d = 2\pi f_d$.

The resultant of **a** and **b** is the vector **c**, and it is the signal corresponding to **c** that is detected by the

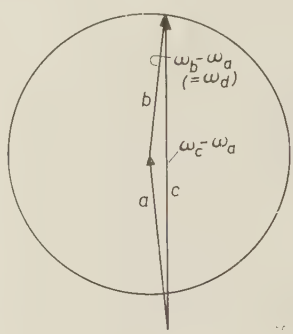


Fig. 3

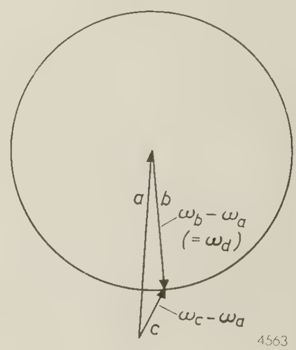


Fig. 4

Figs. 3 and 4. Vectors **a** and **b** represent the intermediate-frequency signals corresponding respectively to direct and indirect reception of an FM signal. Vector **a** is stationary, vector **b** rotates alternately anti-clockwise and clockwise with an angular velocity $\omega_d = 2\pi f_d$ in accordance with eq. (2). The resultant **c** represents the signal to which, after limiting, the discriminator responds. Vector **c** shows amplitude modulation and an irregular angular velocity (ω_c is small at the moment represented in fig. 3; it is large and in the opposite sense to ω_d at the moment to which fig. 4 refers). In both figures $b/a = 0.8$.

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[5]) J. van Slooten, FM reception under conditions of strong interference, Philips tech. Rev. **22**, 352-360, 1960/61 (No. 11).

[6]) In variable reception conditions it may happen that b becomes greater than a. In that case b becomes the desired and a the interfering signal, and in the following considerations, a and b change places. See article cited under [5]).

discriminator. The function of the latter is to deliver an audio-frequency signal whose instantaneous amplitude is proportional to the frequency deviation of the input signal, i.e. in this case proportional to the instantaneous value ω_c of the angular velocity of the vector **c**.

It is easily seen from figs. 3 and 4 that ω_c shows marked variations during one revolution of **b**, particularly if b is not much smaller than a. When **b** points roughly in the opposite direction from **a** (fig. 4), then **c** rotates faster (and in the opposite sense) than when **b** is more or less in line with **a** (fig. 3). There is thus no simple relation between the angular velocity of **c** and that of **b**. The fact that this must cause distortion of the audio signal is evident. We shall presently examine this phenomenon in quantitative terms.

It also appears from figs. 3 and 4 that the vector **c** changes in length during every revolution of **b**, varying from the maximum value $a + b$ to the minimum value $a - b$, which may be zero. The effect of this amplitude modulation will be considered separately.

Distortion due to frequency modulation of **c**

Let ψ be the phase of vector **b**, and φ the phase of vector **c** (*fig. 5*), then

$$\varphi = \tan^{-1} \frac{b \sin \psi}{a + b \cos \psi}.$$

By differentiating φ with respect to time and writing x for the ratio b/a, we find for the angular velocity ω_c of the vector **c**:

$$\omega_c = \frac{x + \cos \psi}{x + x^{-1} + 2 \cos \psi}\, \omega_b , \quad . \quad . \quad (3)$$

where $\omega_b = d\psi/dt$.

To arrive at the instantaneous frequency f_c of the resultant signal, which corresponds to the vector **c**, it must be remembered that as a result of keeping vector **a** stationary we must now add f_a to the frequency of **c**, and also that the calculated angular velocity ω_b corresponds in reality to the difference frequency $f_d = f_a - f_b$. We then find from (3):

$$f_c = f_a + \frac{x + \cos \psi}{x + x^{-1} + 2 \cos \psi}\, f_d .$$

In *fig. 6*, f_c is plotted as a function of ψ for various values of the signal ratio x; the angle ψ runs from zero to 2π (one revolution of **b**) and we assume that in this single period there is no significant change in the frequencies f_a and f_b. At $x = 0$ ($b = 0$, interfering signal absent) f_c is obviously equal to f_a; at $x = 0.25$ a distinct fluctuation occurs in f_c; at $x = 0.5$ the fluctuation is more pronounced, and at $x = 0.75$ it has become a sharp peak at $\psi = \pi$. In the limiting case $x = 1$ (indirect signal just as strong as the direct signal) f_c is always equal to $f + \frac{1}{2}f_d$, except at $\psi = \pi$, where f is discontinuous and equal to $-\infty$. It appears, then, as mentioned above,

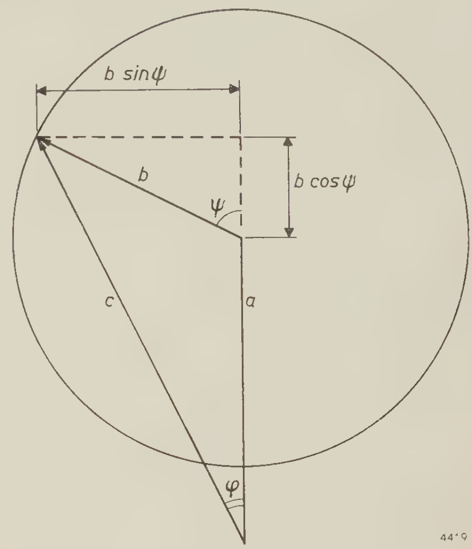

Fig. 5. Illustrating the derivation of the instantaneous frequency f_c and the amplitude modulation of vector **c**.

that the instantaneous frequency f_c of the resultant signal may cover a much wider band than the instantaneous frequency of the constituent signals individually.

The discriminator has to deliver an output signal whose magnitude at any given instant is proportio-

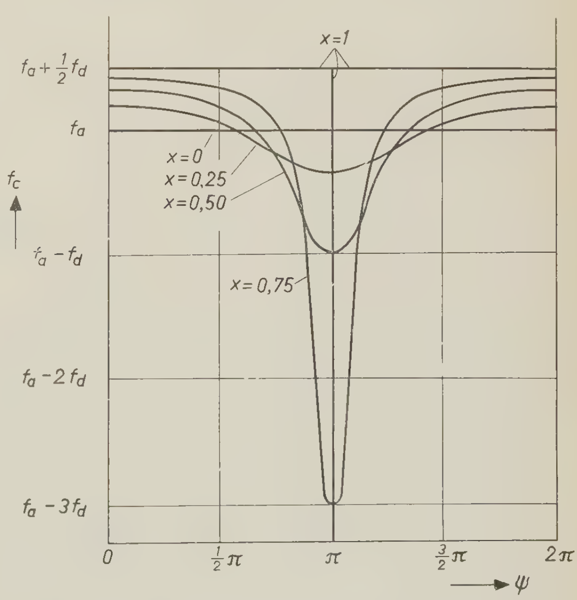

Fig. 6. Instantaneous frequency f_c of vector **c** in figs. 3, 4 and 5, as a function of ψ and for various values of signal ratio $x = b/a$.

nal to the instantaneous frequency deviation of the input signal. For example, if the instantaneous frequency f_c of the input signal has the form shown by the curve for $x = 0.75$ in fig. 6, the output voltage will have a form that corresponds to the fluctuation of f_a (i.e. the original audio signal) except for a superimposed peak at $\psi = \pi$ in each period $2\pi/\omega_b$ of the vector **b**. The output signal will then have an appearance such as that in *fig. 7*.

The number of peaks per period T is equal to the number of complete revolutions of vector **b** (the vector **a** being stationary) in the time T. This number can be determined as follows.

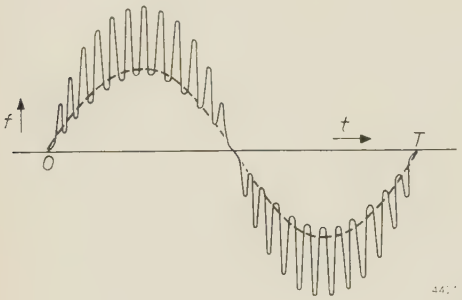

Fig. 7. Broken line: sinusoidal audio signal in undistorted reception. Solid line: with reception via two transmission paths, a peak occurs whenever the vector **c** (fig. 5) has a high angular velocity ($\psi \approx 180°$). Such a waveform occurs when the phase difference of the modulations of the two received signals is 180°.

We represent the phase of the signal that follows the shorter transmission path by

$$\Phi_a = m \sin \Omega t + \omega_0 t$$

and that of the other signal by

$$\Phi_b = m \sin \Omega(t-\tau) + \omega_0(t-\tau)\,.$$

The instantaneous phase difference of the two signals is then:

$$\Phi_d = \Phi_a - \Phi_b = m\,\{\sin \Omega t - \sin \Omega(t-\tau)\} + \omega_0\tau =$$
$$= 2m \sin \tfrac{1}{2}\Omega\tau \cos \Omega(t-\tfrac{1}{2}\tau) + \omega_0\tau\,. \quad \quad (4)$$

This phase difference, then, varies with time in accordance with a cosine function, with the frequency $\Omega/2\pi$. During the quarter period in which the cosine increases from zero to 1, the phase difference changes according to (4) by an amount of $2m \sin \frac{1}{2}\Omega\tau$ radians. In a complete period T the variation therefore amounts to $8m \sin \frac{1}{2}\Omega\tau$ radians. The number of times that the vector **b** in the time T sweeps an angle of 2π radians (giving rise to one peak) is thus $\frac{4}{\pi}\, m \sin \frac{1}{2}\Omega\tau$, and this occurs per second

$$f_r = \frac{4}{\pi}\, \Delta f \sin \tfrac{1}{2}\Omega\tau \text{ times}\,, \quad \quad (5)$$

since $m = T\Delta f$; the quantity $\overline{f_r}$ is the average repetition frequency of the peaks.

Since the phase angle swept per period is generally not an exact multiple of 2π, some peaks will not be completely formed.

Because the frequency difference f_d varies (see fig. 2 or eq. (2)), the repetition frequency f_r of the peaks fluctuates about the mean value $\overline{f_r}$ given by (5). The peaks are farthest apart at the moments when $f_a = f_b$ (i.e. at the points where the two sine waves in fig. 2 intersect) and are closest together midway between these moments. A high pulse rate is associated with a large amplitude. During the time $\frac{1}{2}T$ in which f_b is greater than f_a, the frequency difference f_d is negative and the peaks point upwards; during the other half cycle they point downwards (see fig. 7).

Concerning the numerical value of f_r (eq. (5)) it is difficult to be definite, since the factor $\sin \frac{1}{2}\Omega\tau$ with varying Ω can assume any value between zero and 1. The frequency deviation Δf, which may go up to 75 kc/s, amounts on an average to no more than about 15 kc/s. In the most favourable case ($\sin \frac{1}{2}\Omega\tau = 1$) the value of $\overline{f_r}$ at the average deviation may lie near the threshold of audibility (20 kc/s), and exceeds it only when the deviation increases. The repetition frequency f_r fluctuates, as mentioned, around the mean value $\overline{f_r}$, thereby varying from zero to a value greater than $\overline{f_r}$. Evidently, therefore, f_r can only be above the threshold of audibility during a *part* of the period T. In this part the peaks are generally crowded together. A peak represents a frequency deviation, the mean value of which, $\overline{f_c}$, follows from eq. (3) (if we regard the angular velocity ω_b as constant during the short time in which a peak is formed):

$$\overline{f_c} = \frac{\omega_c}{2\pi} = \frac{\omega_b}{2\pi^2} \int_0^{\pi} \frac{x + \cos\psi}{x + x^{-1} + 2\cos\psi}\, d\psi\,. \quad \quad (6)$$

It can be shown that for x smaller than unity this integral is zero. This means that the peaks in the time interval in question are practically inaudible.

What is the situation in practice? Apart from on the transit-time difference τ, the phase difference depends on the audio frequency p, for at a low audio frequency the phase varies less per unit time (and thus also in the time τ) than at a high audio frequency. Since, in the spectrum of music and speech, numerous frequencies occur which are fairly uniformly distributed over the audio range, there will always be a great many phase differences, some of them small. This, together with the fact that the average frequency deviation is only about 15 kc/s,

accounts for the presence of an audible interference in music and speech whenever the strength ratio x of the signals is relatively large.

In the foregoing we have tacitly assumed an ideal discriminator, i.e. one whose bandwidth is unlimited. In reality the bandwidth of a discriminator is limited, and if it is not large enough to deal completely with the peaks of f_c, the situation is even more unfavourable than described. The integral in (6) is then no longer equal to zero, so that every clipped peak gives rise to an audible component in the audio signal. Since the peaks representing the largest frequency deviations suffer most from the limited bandwidth, distortion occurs in the audio signal even when f_r is above the audio limit.

To prevent this, it is necessary to ensure that the bandwidth B of the non-linear portion of the receiver — i.e. the discriminator and the preceding limiter or limiters — is above a certain minimum. This minimum follows from the fact that the maximum frequency at the top of the peak is:

$$\frac{b}{c}(f_{\mathrm{d}})_{\max} + (f_{\mathrm{a}})_{\max} = \frac{b}{a-b} \times 2\Delta f + \Delta f = \frac{1+x}{1-x}\Delta f.$$

The required bandwidth is twice as large, i.e.

$$B \geq \frac{1+x}{1-x} \cdot 2\Delta f. \quad . \; . \; . \; . \quad (7)$$

By solving (7) for x, we find the signal ratio x at which a given bandwidth B can only just deal with the peaks:

$$x = \frac{B - 2\Delta f}{B + 2\Delta f}.$$

In good FM receivers 400 kc/s is a usual bandwidth for the limiter and discriminator section. At a deviation Δf of 15 kc/s, $x = 0.86$ is the value at which the peaks can still just be handled and at which the additional distortion is just suppressed. At a deviation of 75 kc/s the limit lies at $x = 0.455$. Where x is in the neighbourhood of 1 the bandwidth has to be very large indeed, e.g. almost 6 Mc/s at $x = 0.95$ (and $\Delta f = 75$ kc/s).

Distortion due to amplitude modulation of **c**

The foregoing section related to the part contributed to the distortion by the irregular angular velocity of the vector **c** (fig. 5). A second contribution is due to the variations in the magnitude of the vector **c**, i.e. to the amplitude modulation (AM) present in the signal corresponding to **c**.

It can be seen from fig. 5 that

$$c = \sqrt{a^2 + b^2 + ab \cos \psi}\,.$$

In *fig. 8* the variation of the amplitude c as a function of ψ is plotted for a single revolution of the vector **b**, for different values of the signal ratio x. Each of these curves has a modulation depth equal to x.

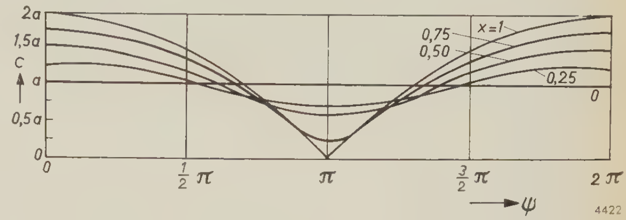

Fig. 8. Amplitude modulation of vector **c** (fig. 5). The variation of the amplitude c is shown as a function of the angle ψ between **a** and **b**, for various values of the signal ratio $x = b/a$.

It is the task of the limiter to suppress this amplitude modulation in order that the output signal from the discriminator shall not have a component possessing the distortion shown in fig. 8. FM broadcast receivers usually have two limiters. The function of the first limiter is performed by the last valve in the intermediate-frequency amplifier; this only acts as a limiter — due to the flow of grid current — when the signal reaches a certain strength. The second limiter — which operates also on weaker signals — is the discriminator itself, if the latter is a ratio detector [7]). The limiting action here is based on the variable damping effect, dependent on the signal amplitude, which a diode exerts on a resonant circuit. Obviously, this damping cannot go lower than zero. If the signal amplitude changes still further in the same direction, the limiting action fails, the diodes being temporarily cut-off. This effect will certainly appear at modulation depths of about 0.7 upwards.

If the signal is strong, the first limiter provides at the least for a 10-fold reduction of the modulation depth; there is then no chance of the detector failing, and the amplitude modulation is effectively neutralized. Where the signal is weak, however, the first limiter is inoperative, and if x reaches the value at which the detector begins to fail, strong AM distortion results.

It should be noted that the fundamental frequency of the AM component is determined by the angular velocity of the vector **b**. As regards the audibility of an insufficiently suppressed AM component, the same remarks apply as to the peaks due to the frequency modulation of the vector **c**.

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[7]) The operation of the ratio detector as a discriminator and limiter is briefly explained in Philips tech. Rev. **17**, 346, 1955/56, and treated in more detail by F. E. Terman, Electronic and radio engineering, McGraw-Hill, New York 1955, 4th impression, p. 610 *et seq.*

The great importance of good AM suppression may be understood by considering the case where AM is not suppressed at all. The low-frequency voltage generated in the discriminator by the *amplitude* modulation of **c** is then found to be roughly $15x$ times higher than the undistorted signal that the discriminator should give (for a frequency deviation of 15 kc/s). Even if x is only 0.1, the AM interference is still 1.5 times stronger than the desired audio signal. On the other hand, where the receiver bandwidth is adequate and the values of x are small, but otherwise under the most unfavourable conditions, the interference due to *frequency* modulation of **c** is only $0.2x$ times the undistorted audio signal. To reduce the total interference it is therefore of paramount importance to have adequate AM suppression.

Measures for reducing FM distortion due to multipath transmission

From the fact that the distortion discussed increases in severity the closer the signal ratio $x = b/a$ approaches unity, it follows that our first countermeasure must be to try to reduce x. For this purpose an aerial having a sharp directional effect should be used, positioned in such a way that x is minimum.

The measures to be taken in the receiver itself consist, as follows from the above considerations, in giving the limiter and discriminator section a large bandwidth, in ensuring rigorous limiting, and in designing a discriminator capable of handling signal ratios close to unity.

This article is not concerned with the means by which these requirements can be met. In the next section, however, we shall describe a signal generator designed to simulate the aerial signal resulting from multipath transmission. With this signal generator it is possible to study the effect of the above-mentioned measures to reduce distortion, even in places where there are no suitable reflecting obstacles in the neighbourhood.

A signal generator for simulating multipath transmission effects in FM reception

The investigators cited in footnotes [1]) and [3]) studied these effects experimentally as well as theoretically. They conducted the radio-frequency FM signal, produced in a generator, along two paths to the receiver under investigation: along a short direct path and along a path having an appreciable delay time. For this second path Corrington used a coaxial cable which had to be more than 3 km long to give a transit time of 16 μsec — corresponding to a detour in the "ether" of less than 5 km. Arguimbau and Granlund, by piezo-electric means, first converted the radio-frequency signal into an ultrasonic vibration; they passed this through a mercury column, at the end of which they converted it back again into an electrical signal. Neither of these methods is convenient if the apparatus is required to be compact and portable. For these reasons we have adopted a different approach to the problem.

Two radio-frequency signals have to be generated, namely a "direct" signal of instantaneous frequency F_a:

$$F_a = F_0 + \Delta f \sin \Omega t ,$$

and a "reflected" signal of instantaneous frequency F_b:

$$F_b = F_0 + \Delta f \sin \Omega(t-\tau) .$$

Here F_0 is the centre frequency, and there is only one audio frequency ($= \Omega/2\pi$). Introducing an angle α which satisfies

$$\alpha = |\Omega t - 2\pi n| < 2\pi ,$$

where n is an integer, we can write the formula for F_b as

$$F_b = F_0 + \Delta f \sin (\Omega t - \alpha) .$$

The two radio-frequency signals can thus in principle be obtained by using the same audio signal to frequency-modulate two signal generators having the same centre frequency F_0, the "delayed" signal being made to lag the "direct" signal by a phase angle α. The latter is produced by a phase shifter which will presently be discussed.

In actual FM reception the transit-time difference τ gives rise not only to a phase difference α between the modulations, but also to a phase difference $\omega_0 \tau$ between the radio-frequency signals; see eq. (4). In order to simulate the latter phase difference, use might be made of a second phase shifter, now for high frequencies. We have not done this, however, since the only effect of the phase angle $\omega_0 \tau$ consists in a displacement of the peaks in fig. 7 over less than the width of one peak. This displacement is of no importance to the study of multipath distortion.

With two separate oscillators it is not possible to satisfy the condition that the two modulated signals shall have the same centre frequency. One common oscillator must therefore be used. Frequency modulation, however, can only be introduced in the oscillator itself; modulating this (in frequency) by two different audio signals would be no use whatsoever, for it would not yield the required two radio-frequency signals each modulated by one of the audio signals.

A system free from this limitation is phase modulation, for here the modulation can be applied *after* the oscillator. Phase modulation was therefore the system we decided to adopt. A phase-modulated signal, however, differs from a frequency-modulated signal in that the frequency deviation is proportional not only to the amplitude of the audio signal but also to the audio frequency. In order to make the result of phase modulation identical with that of frequency modulation, the audio signal is passed through an integrating network; this delivers an output voltage which is inversely proportional to the audio frequency.

Block diagram

The block diagram of the signal generator is shown in *fig. 9*. The oscillator O_1, whose frequency is controlled by a quartz crystal, gives an output having a constant frequency of 3 Mc/s. In the phase modulators PM_a and PM_b the oscillator output is modulated in phase by an audio signal which is applied to the modulators through the above-mentioned integrating networks, I_a and I_b. Two buffers B_{a1} and B_{b1}, each consisting of a simple pentode stage, prevent feedback from the phase modulators to the oscillator.

Fig. 9 refers to the case of a *sinusoidal* audio signal. This is delivered by the signal generator G, and a phase shifter P preceding the integrator I_b provides the variable phase angle α. We shall return in a moment to the circuit arrangements to permit modulation by music and speech.

Following the normal practice in FM transmitters, the maximum frequency deviation is brought to the required 75 kc/s by frequency multiplication. This is done in the stages Mu_{a1}, Mu_{a2}, Mu_{b1} and Mu_{b2}.

For this purpose the total multiplication factor needed is more than 400. A factor of about 30, however, is sufficient to raise the frequency of O_1 (3 Mc/s) to a value within the FM broadcast band (87.5-100 Mc/s). For this reason 18-fold multiplication is applied in Mu_{a1} and Mu_{b1}, after which, by mixing with an auxiliary frequency of 50 Mc/s, the centre frequency is reduced from 54 to 4 Mc/s, the 18-fold frequency deviation being retained. In Mu_{a2} and Mu_{b2} a 24-fold multiplication is then applied, which raises the centre frequency from 4 to 96 Mc/s and brings the total multiplication of the deviation to $18 \times 24 = 432$.

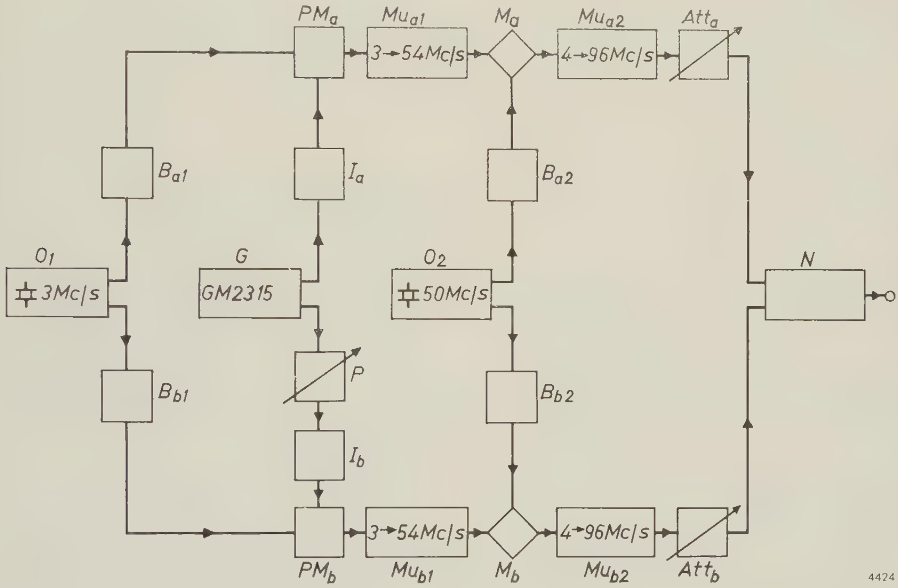


Fig. 9. Block diagram of apparatus for simulating FM multipath reception. Subscripts *a* relate to the "direct" signal, subscripts *b* to the "reflected" signal.
O_1 3 Mc/s crystal oscillator. B_{a1}, B_{b1} buffer stages. G signal generator (type GM 2315). P phase shifter. I_a, I_b integrating networks. PM_a, PM_b phase modulators. Mu_{a1}, Mu_{b1} frequency multipliers (18×). O_2 50 Mc/s crystal oscillator. B_{a2}, B_{b2} buffer stages. M_a, M_b mixing stages. Mu_{a2}, Mu_{b2} frequency multipliers (24×). Att_a, Att_b continuously variable "ladder" attenuators. N matching network. The aerial terminals of the receiver are connected to the output of N.

The auxiliary frequency of 50 Mc/s is generated by the crystal oscillator O_2, and mixing is done in the stages M_a and M_b. The buffers (pentode stages) B_{a2} and B_{b2} prevent undesired coupling between the channels via O_2.

The mixing process would not be necessary if the frequency of O_1 were low enough (e.g. 220 kc/s) to allow the same high multiplication factor used for the frequency deviation to be applied for the central frequency. A frequency as low as this for O_1, however, would have entailed considerable difficulties with the bandwidth.

At the outputs of the multipliers Mu_{a2} and Mu_{b2} a continuously variable attenuator is connected (Att_a and Att_b, respectively); this is a so-called "ladder" attenuator, which has the property that the impedance remains constant (here 50 ohms), whatever the attenuation. Through coaxial cables, whose characteristic impedance is likewise 50 ohms, both signals are conducted to a matching network N, which terminates the cables with 50 ohms and shows the same impedance at its output. Here the terminals of the receiver are connected.

The circuit shown in *fig. 10* makes it possible to shift the sinusoidal modulation of the "reflected" signal by a variable phase angle α without changing the amplitude of the audio signal. When the resistance R is raised from zero to R_{max}, the phase difference α between the output and input voltages increases from zero to almost π, provided that ΩCR_{max} is large compared with unity and the output terminals are not loaded. In order to vary α from zero to almost 2π, two of these circuits are connected in cascade. The phase modulators will be dealt with in the next section.

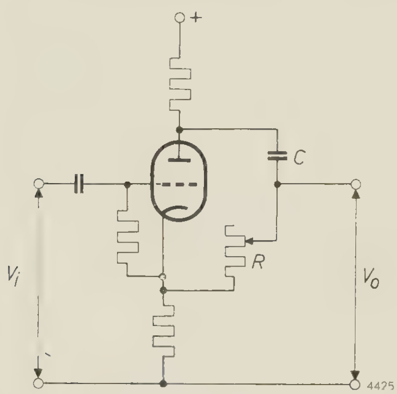


Fig. 10. One of the two cascade stages forming the phase shifter. When the resistance R is varied from zero to R_{max}, the phase angle α between output voltage V_o and input voltage V_i increases from zero to 2 $\tan^{-1} \Omega CR_{max}$, that is, to nearly π if ΩCR_{max} is large compared with unity. Since there are two stages in cascade, the total phase shift runs from zero to nearly 2π. The magnitude of V_o is not thereby affected.

A photograph of the equipment is shown in *fig. 11*. Two oscillograms obtained with it can be seen in *fig. 12*. The oscillogram in fig. 12*a* (analogous to fig. 7) relates to a receiver with a good AM limiter; that of fig. 12*b* refers to a receiver with poor AM limitation. In the second case the interference was much more audible than in the first.

The phase modulators

For phase modulation a network is needed where the change of the phase difference between the output and input voltage is proportional to the instantaneous amplitude of the audio signal. A network with this property is a bandpass filter consisting of two coupled LC circuits in which the capacitances C are voltage-dependent. When the capacitances are varied, the tuning of the bandpass filter changes accordingly, and so therefore does the phase Θ of the output voltage. *Fig. 13* shows the variation of Θ as a function of βQ for the case of a critically-coupled bandpass filter of identical primary and secondary Q. Here β is the relative detuning, $\approx 2(f_{res} - f)/f$ (where f is the signal frequency and f_{res} is the resonant frequency of the filter). As can be seen, the response characteristic is virtually straight from $\beta Q = -1.7$ to $+1.7$ (Θ from -100 to $+100°$). If Q has the easily achievable value of 85, for example, then β can remain small. In this case, then, the phase modulation is practically linear if β does not exceed the value $1.7/85 = 0.02$. For such slight detuning, β is approximately equal to $\Delta C/C$, so that the capacitance variation ΔC may amount to a maximum of about 2% of C.

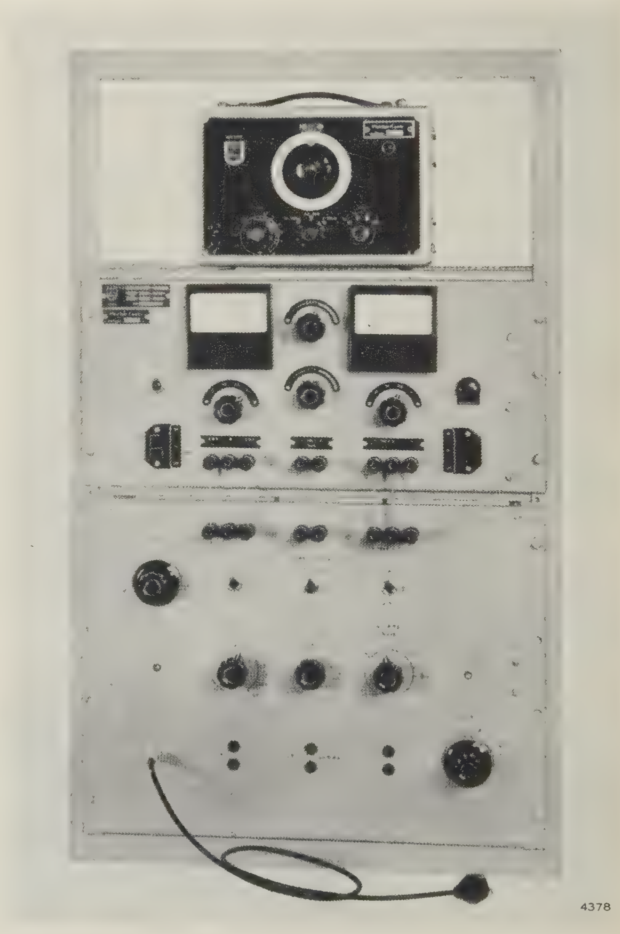
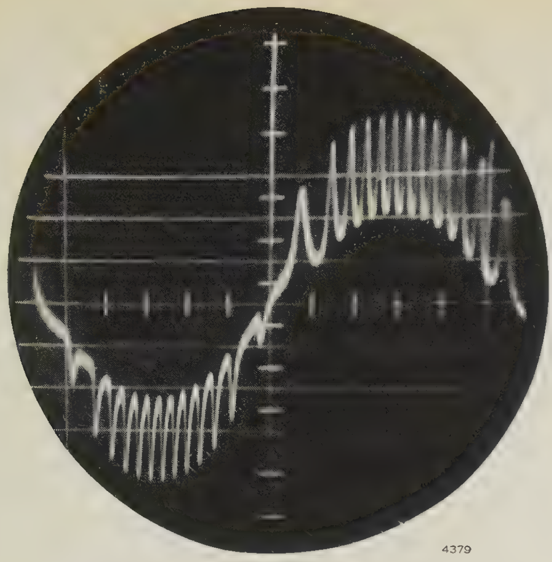
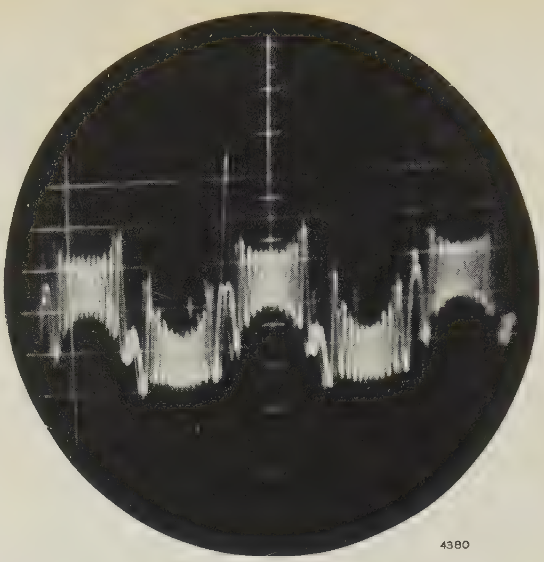


Fig. 11. The complete simulator. Above, the signal generator type GM 2315 (which can be replaced by a delay device with magnetic recording; see fig. 15). The panel below it is the power pack, and the bottom panel is the actual apparatus.



a



b

Fig. 12. Oscillograms of the output voltage from the discriminator in an FM receiver connected to the simulator described. The modulation was sinusoidal. *a*) Receiver with satisfactory AM limiting (cf. fig. 7). *b*) Receiver with inadequate AM limiting.

Semiconductor diodes in the cut-off state behave as voltage-dependent capacitances. The capacitance C depends in the following way on the applied reverse voltage V:

$$C = \frac{K_1}{\sqrt{-K_2-V}},$$

where K_1 and K_2 are diode constants (K_2 is nega-

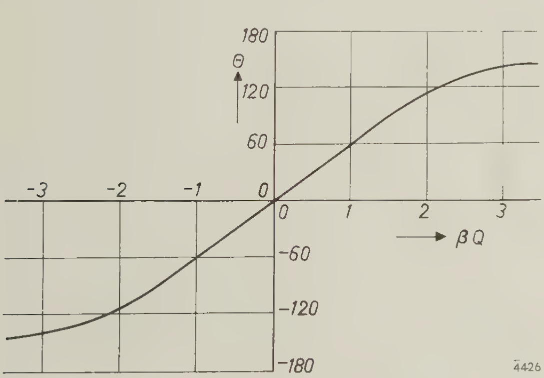


Fig. 13. Phase variation Θ of the output voltage from a critically-coupled bandpass filter having an identical primary and secondary Q factor, as a function of Q times the relative detuning β.

tive). *Fig. 14* shows the variation of C with voltage V measured on a silicon diode.

If a small alternating voltage is superposed on the direct voltage V, the change of C is roughly proportional to the alternating voltage. This principle is applied in the phase modulators PM_a and PM_b (fig. 9).

Modulation by music or speech

In order to make the distorted output signal from the receiver visible in an oscillogram, the obvious method is to modulate with a sinusoidal audio signal; this case is represented in the block diagram in fig. 9.

After the signal generator had been thus designed, however, the need arose for some means of judging by ear the quality of received music or speech when the distortion described is present. For this purpose the apparatus was extended with a device for modulating by music or speech. The signal generator and the phase shifter are then put out of action — the latter because the phase angle α it delivers is not proportional to the audio frequency, which means

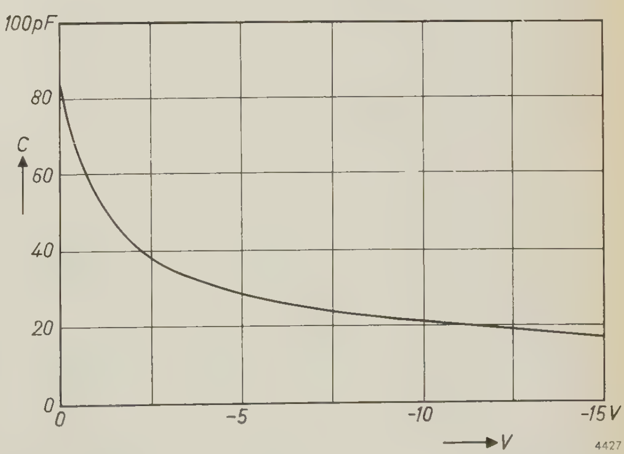


Fig. 14. Measured capacitance C of a silicon diode as a function of the DC reverse voltage V.

that the time delay τ would be dependent on the audio frequency. To obtain a variable and frequency-independent time delay, we record the speech or music on a magnetic tape. When the apparatus is working, the tape passes over two playback heads spaced a distance s apart. This distance, which corresponds to the desired time delay τ, must be variable (at normal tape speed) from 0 to about 0.2 mm; s is therefore always much smaller than the dimensions of the playback heads, and this calls for a special design.

A design which satisfactorily meets the requirements is illustrated schematically in *fig. 15*. The audio signal is recorded over almost the entire width of the tape, and the two playback heads each scan only one half of the width. The heads are mounted one above the other; the lower head is stationary, and the upper head is rotatable in relation to the other about a common axis. The distance s between the gaps of the heads varies with the angle of rotation. This distance is increased in a fixed ratio by levers H_1 and H_2, and can be read from the micrometer M.

At a tape speed of 19 cm/sec, a time delay up to 1 millisecond can be achieved in this way, which simulates a difference of 300 km in the length of the radio transmission paths. This is in fact more than in the cases ever encountered in practice. The distance s can be adjusted to an accuracy of 2 μ, which corresponds to a path-length difference of about 3 km. Small differences in path-length can thus equally well be simulated.

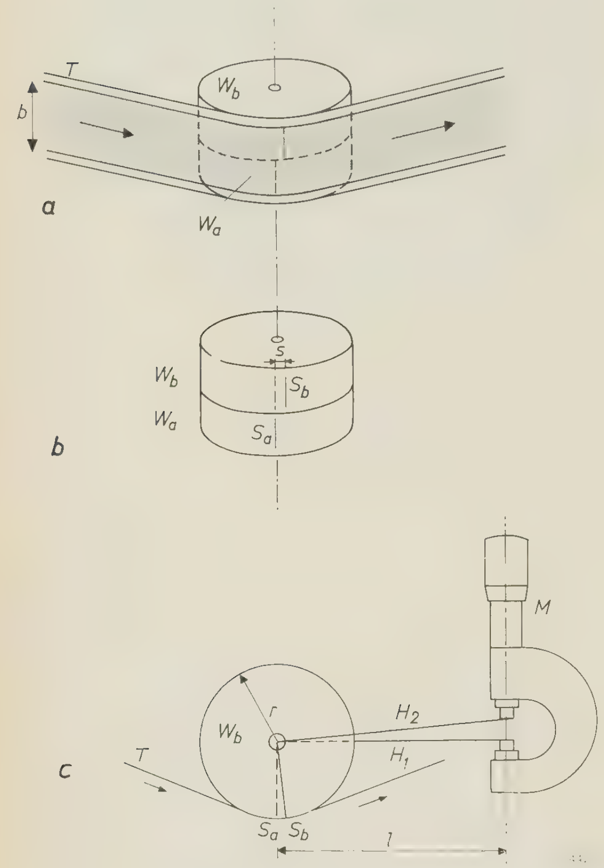


Fig. 15. Scanning of a magnetic tape by two playback heads, the distance between the gaps of which is smaller than the dimensions of the heads.
a) On the tape T music or speech is recorded over the full track width b.
b) The playback heads W_a and W_b each scan half the width of the track (the tape itself is not shown). The head W_b is turned in relation to the fixed head W_a, so that the gap S_b is passed somewhat later than the gap S_a. The separation s corresponds to the time delay τ between the signals of W_a and W_b, and amounts to less than 0.2 mm.
c) The levers H_1 and H_2 effectively increase the separation s by the (fixed) ratio l/r to a value that can be measured accurately by the micrometer M.

Summary. In mountainous regions, two or more FM transmission paths of different length may exist between the transmitting and receiving aerials, as a result of reflections from mountain ridges. This can cause distortion in reception, particularly if the received signals are of roughly the same strength. The discriminator then receives the resultant of the signals, and the vector representing this resultant exhibits an irregular angular velocity, which is not directly related to the modulation of the transmitter. Furthermore, the resultant is subject to amplitude modulation, which also causes distortion. Means of improving reception are the use of a sharply directional aerial, increasing the bandwidth of the limiter and discriminator, rigorous limiting, and the use of a discriminator capable of handling a signal ratio close to unity.

In order to simulate the effects in the laboratory, irrespective of terrain or conditions of reception, an apparatus has been designed which delivers two RF signals, one delayed and the other not, which are modulated in frequency by an audio signal (sine wave; music or speech). The time delay is continuously variable, and simulates a maximum path difference of 300 km.

CIRCULAR OPTICAL ABSORPTION WEDGES

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The level of illumination in nature increases from about 0.5 lux at dusk to 50 000 lux in bright sunlight, i.e. by a factor of 10^5. For outdoor work a television camera should preferably be fitted with a camera tube sensitive enough to respond to the lower of these two levels. In stronger illumination the iris diaphragm in the camera lens must then be stopped down considerably, as the camera tube has a latitude of only about a factor 10 in the average intensity of illumination it can handle. If the aper-

ture can be varied with the diaphragm between, say, *f*.2 and *f*.20, this gives an attenuation factor of 100 at the most. In order to use the camera at the higher of the illumination levels mentioned, additional attenuation by a factor of 100 is therefore needed.

This attenuation can be achieved using a neutral absorption filter. However, as every amateur photographer knows, the iris diaphragm of a camera serves not only to regulate the light entering the lens but also to regulate the depth of focus. If an extra attenuating element is to be introduced in the camera, it too should be variable (preferably continuously), so that the depth of focus can be independently selected in as wide a range of illumination levels as possible. This also applies where automatic mechanisms are concerned. A continuously variable attenuating element can be made by designing the absorption filter in the form of a movable density wedge.

We have devised a method of making absorption wedges in the form of a circular disk, the density varying with azimuth. This enables the light transmission to be varied by rotating the disk. It is not practicable to make such a circular wedge by grinding absorbent glass to a continuously tapering thickness — the method often adopted for straight wedges. Our method is therefore to effect the absorption through a non-scattering layer of material vapour-deposited on a glass disk, the thickness of the layer being given the azimuthal density variation required (if necessary non-linear). The set-up used for this purpose is illustrated in *fig. 1*.

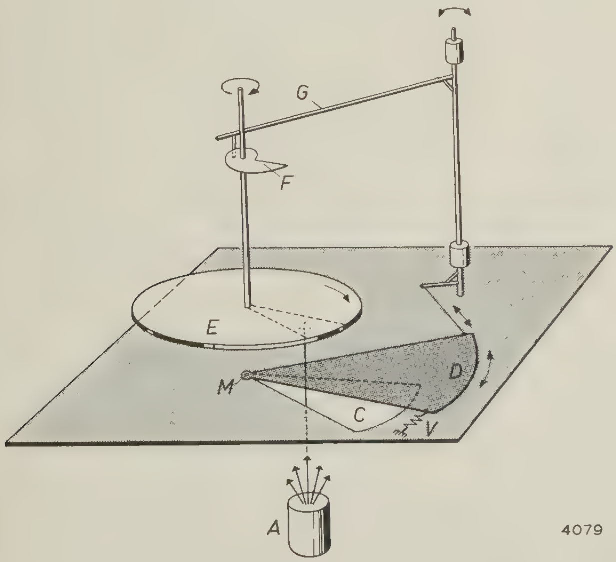

Fig. 1.

The source *A* of the material to be vapour-deposited is situated below a fixed plate in which an aperture *C* is cut in the form of a sector. The width of the aperture can be varied by means of a plate *D* which pivots about the point *M*. Mounted above this assembly is the round glass plate *E* on which the layer is to be deposited and which is rotated at a uniform speed about a vertical spindle through *M*.

Fig. 2.

Fixed to this spindle is a cam *F* which, in turning, displaces a lever *G* and thereby moves the cover plate *D* against the action of a spring *V*. Thus, in every azimuthal position of the glass plate a sector of a certain size is exposed to the vaporized material; the density law of the circular wedge is thus governed by the shape of the cam. The deposition of the layer can be continued over any arbitrary number of complete revolutions of the glass plate, so that — provided the source *A* operates constantly — any desired maximum attenuation can be achieved.

Suitable materials for vapour-deposition are metals or mixtures of metals with SiO; various organic dyes may also be used.

Fig. 2 shows a photograph, taken in transmitted light, of a circular absorption wedge made by the method described. Since the deepest black and the brightest white obtainable with conventional reproduction techniques have a brightness ratio of no better than 15 to 20, it is not possible to do full justice here to the total range of transmission (factor 100) covered by this wedge.

J. van der WAL.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS BY THE STAFF OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN

Reprints of these papers not marked with an asterisk * can be obtained free of charge upon application to Philips Electrical Ltd., Centry House, London W.C. 2, where a limited number of reprints are available for distribution.

2825: H. Bremmer: Some theoretical investigations on fading phenomena (Statistical methods in radio wave propagation, Proc. Symp. Univ. Calif., Los Angeles, June 1958, edited by W. C. Hoffman, pp. 37-39, Pergamon, London 1960).

The author first discusses a statistically-fluctuating signal $h(t) = A(t) \cos \{\omega_0 t + \varphi(t)\}$, where A and φ are slowly varying functions of time t. Under very general assumptions it possesses the property that N_A, the average number of times the amplitude passes through its median value per unit time interval, is about three times greater than N_φ, the average number of crossings of the phase through any special value. If such a fluctuating signal is superposed on a much larger constant signal of fixed amplitude, frequency and phase, it is shown that $N_A \approx N_\varphi$. The author considers the application of these results to the fading of radio signals due to tropospheric turbulence effects.

2826: Th. G. Schut and W. J. Oosterkamp: Die Anwendung elektronischer Gedächtnisse in der Radiologie (Elektron. Rdsch. 14, 19-20, 1960, No. 1). (The application of electronics to radiology; in German.)

The information presented in a fluoroscopic image can be retained by the eye for only about 0.1 sec. Full observation of such an image, however, calls for a much longer time, e.g. 10 sec. If it is possible to store the momentary image in some form of "memory" and subsequently make it visible for a sufficient length of time, the X-ray dose to the patient can be considerably reduced. The radiograph is one such memory, but has the drawback of not being immediately available. The authors discuss other methods that overcome this drawback, in particular the recording of X-ray images on a magnetic wheel store. See also Philips tech. Rev. 22, 1-10, 1960/61 (No. 1).

2827: L. A. Æ. Sluyterman and J. M. Kwestroo-Van den Bosch: Sulphation of insulin and electrophoresis of the products obtained (Biochim. biophys. Acta **38**, 102-113, 1960, No. 1).

In connexion with investigations concerning the chemical modification of proteins, the SO_3 complexes of a few tertiary amines were tested for their ability to introduce SO_3 groups into insulin. Pyridinium sulphonic acid was found to be the most suitable one. By variation of the reaction conditions, insulin preparations of various sulphate content were prepared and subjected to paper electrophoresis at pH 1.7. A total number of 13 well defined, approximately equidistant bands could be observed, corresponding to insulin molecules carrying different electrical charges and covering a range from +6 units (native insulin) to —6 units (completely sulphated insulin). The biological activity of the preparations decreased with increasing sulphate content.

2828: J. S. C. Wessels: Photoreduction of 2,4-dinitrophenol by chloroplasts (Biochim. biophys. Acta **38**, 195-196, 1960, No. 1).

The author had formerly found (see Abstract No. **2776**) that 2,4-dinitrophenol (DNP) is able to catalyse the synthesis of adenosine triphosphate (ATP) by spinach chloroplasts. He now reports that illuminated chloroplasts of spinach are capable of reducing DNP to 2-amino-4-nitrophenol, and that the latter compound can serve as a co-factor of photosynthetic phosphorylation.

2829: L. A. Æ. Sluyterman: The effect of oxygen upon the micro-determination of histidine with the aid of the Pauly reaction (Biochim. biophys. Acta **38**, 218-221, 1960, No. 2).

The colour obtained upon the addition of diazosulphanilic acid to histidine in alkaline medium (Pauly reaction) is bleached rather suddenly after a certain lag. This lag is shorter the more oxygen is present in the alkaline reaction medium.

A method of determining histidine on a micro scale, consisting of an improved Pauly reaction after paper-chromatographic separation, is described in detail.

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